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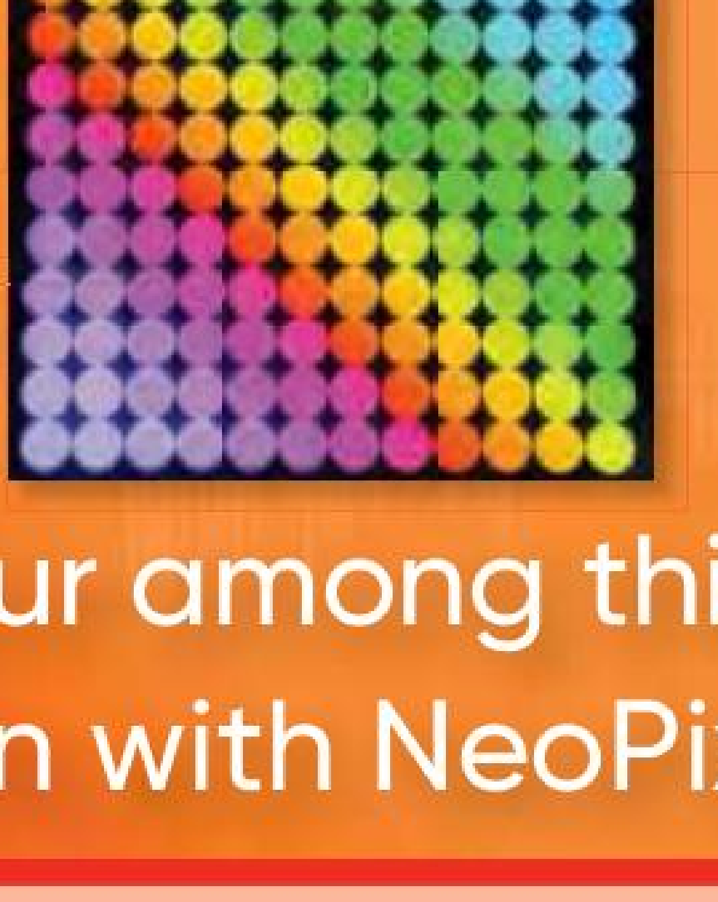

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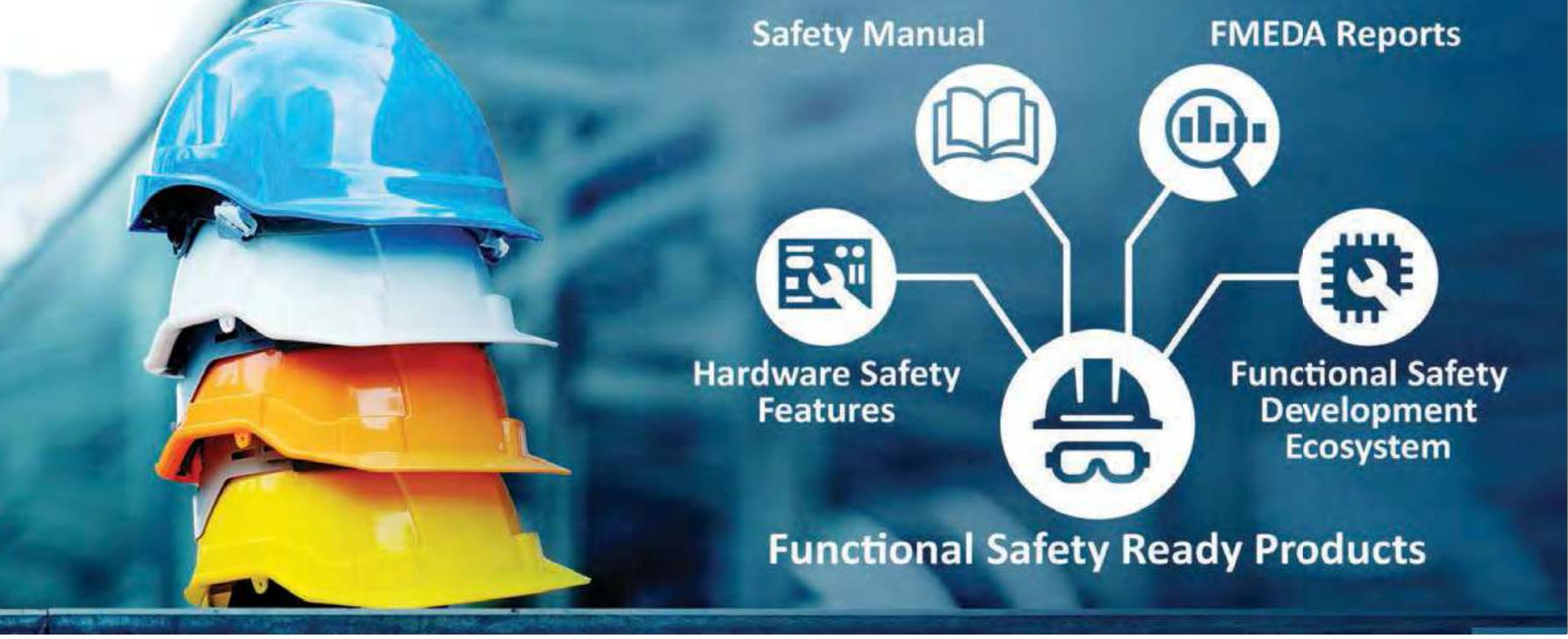
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Practical Electronics

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October 2020
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The November 2020 issue of *Practical Electronics* will be published on Thursday, 1 October 2020 – see page 72.

WIRELESS FOR THE WARRIOR

by LOUIS MEULSTEE

THE DEFINITIVE TECHNICAL HISTORY OF RADIO COMMUNICATION EQUIPMENT IN THE BRITISH ARMY

The *Wireless for the Warrior* books are a source of reference for the history and development of radio communication equipment used by the British Army from the very early days of wireless up to the 1960s.

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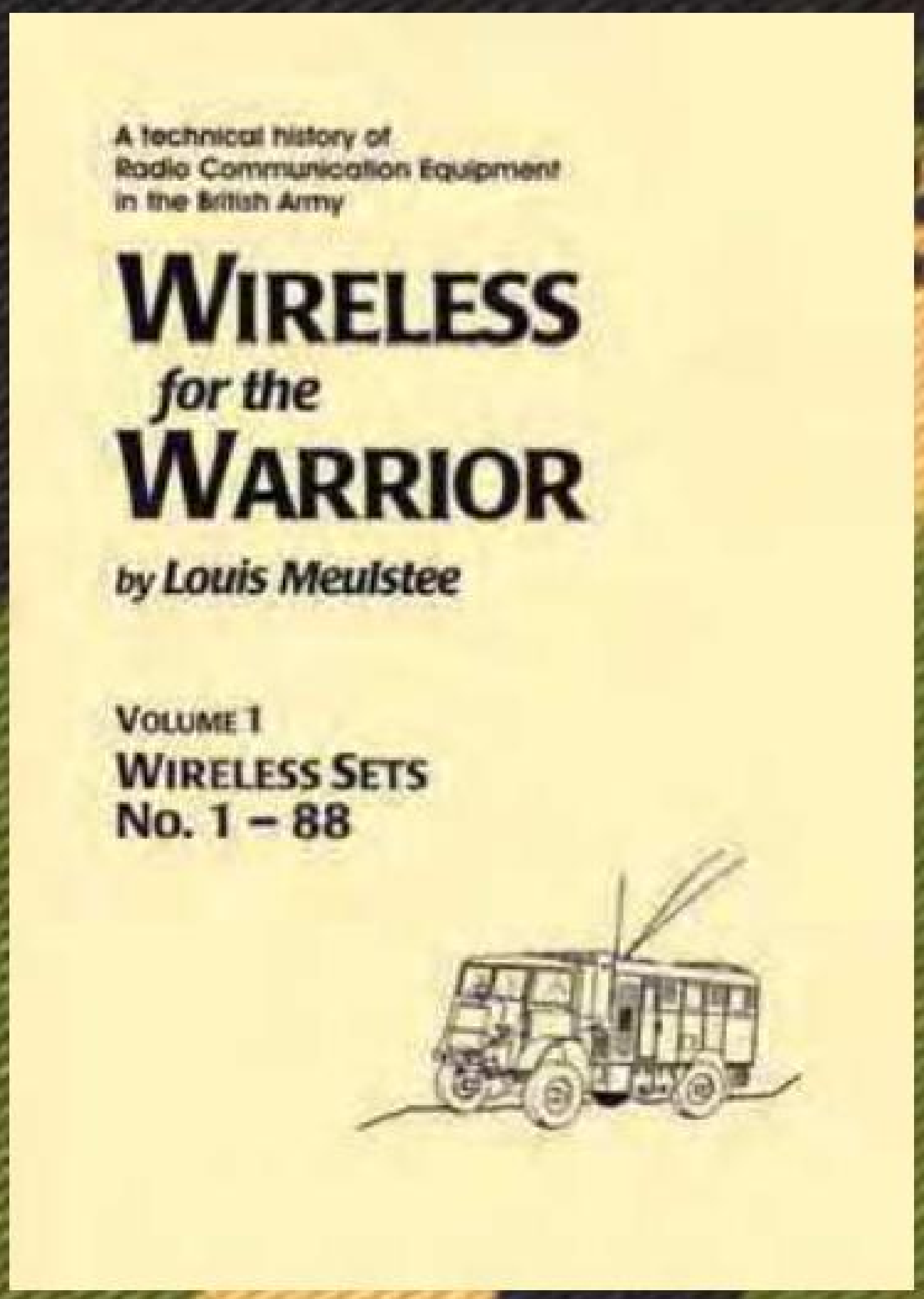
Volume 1 and *Volume 2* cover transmitters and transceivers used between 1932-1948. An era that starts with positive steps taken to formulate and develop a new series of wireless sets that offered great improvements over obsolete World War I pattern equipment. The other end of this timeframe saw the introduction of VHF FM and hermetically sealed equipment.

Volume 3 covers army receivers from 1932 to the late 1960s. The book not only describes receivers specifically designed for the British Army, but also the Royal Navy and RAF. Also covered: special receivers, direction finding receivers, Canadian and Australian Army receivers, commercial receivers adopted by the Army, and Army Welfare broadcast receivers.

Volume 4 covers clandestine, agent or 'spy' radio equipment, sets which were used by special forces, partisans, resistance, 'stay behind' organisations, Australian Coast Watchers and the diplomatic service. Plus, selected associated power sources, RDF and intercept receivers, bugs and radar beacons.



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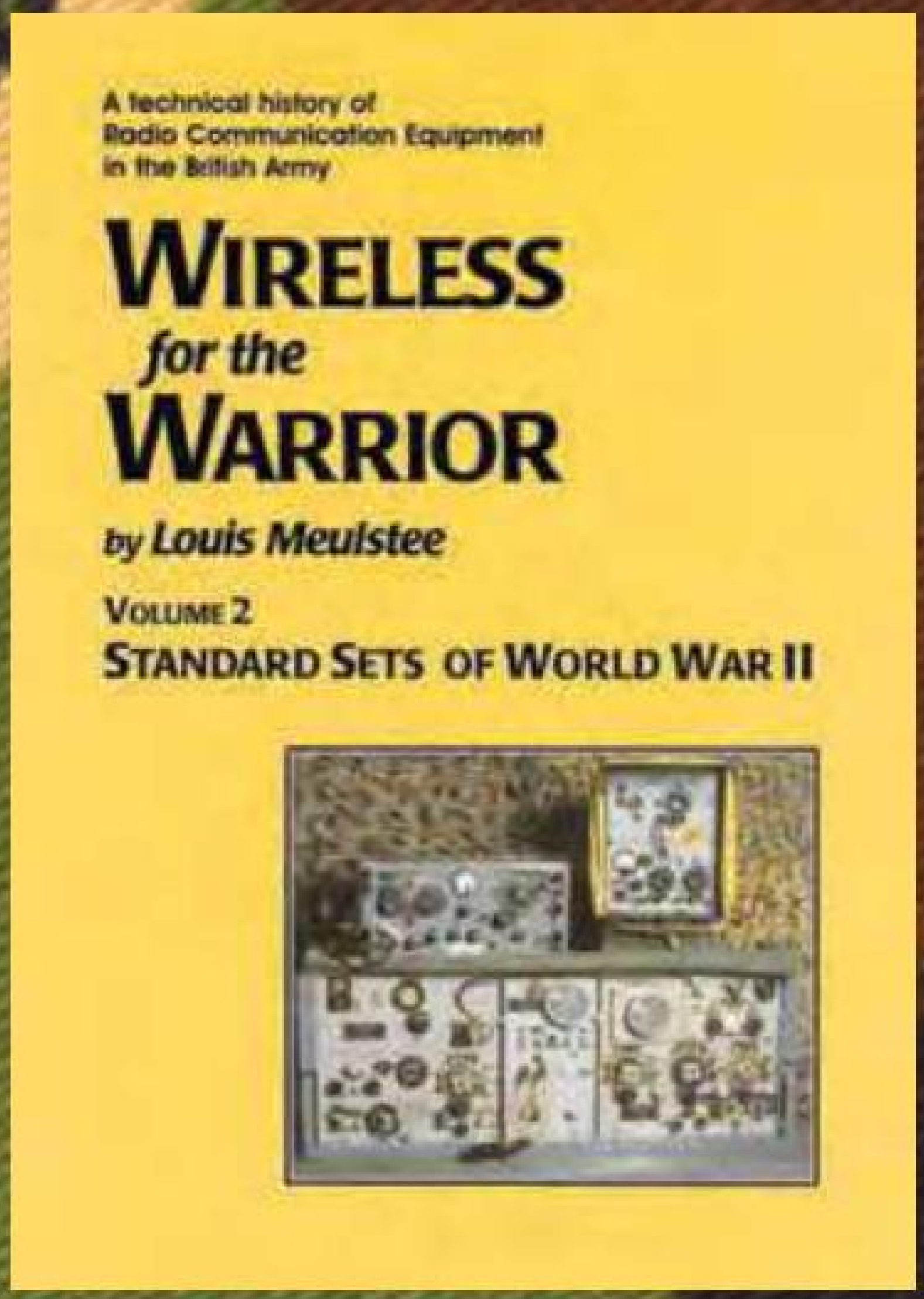

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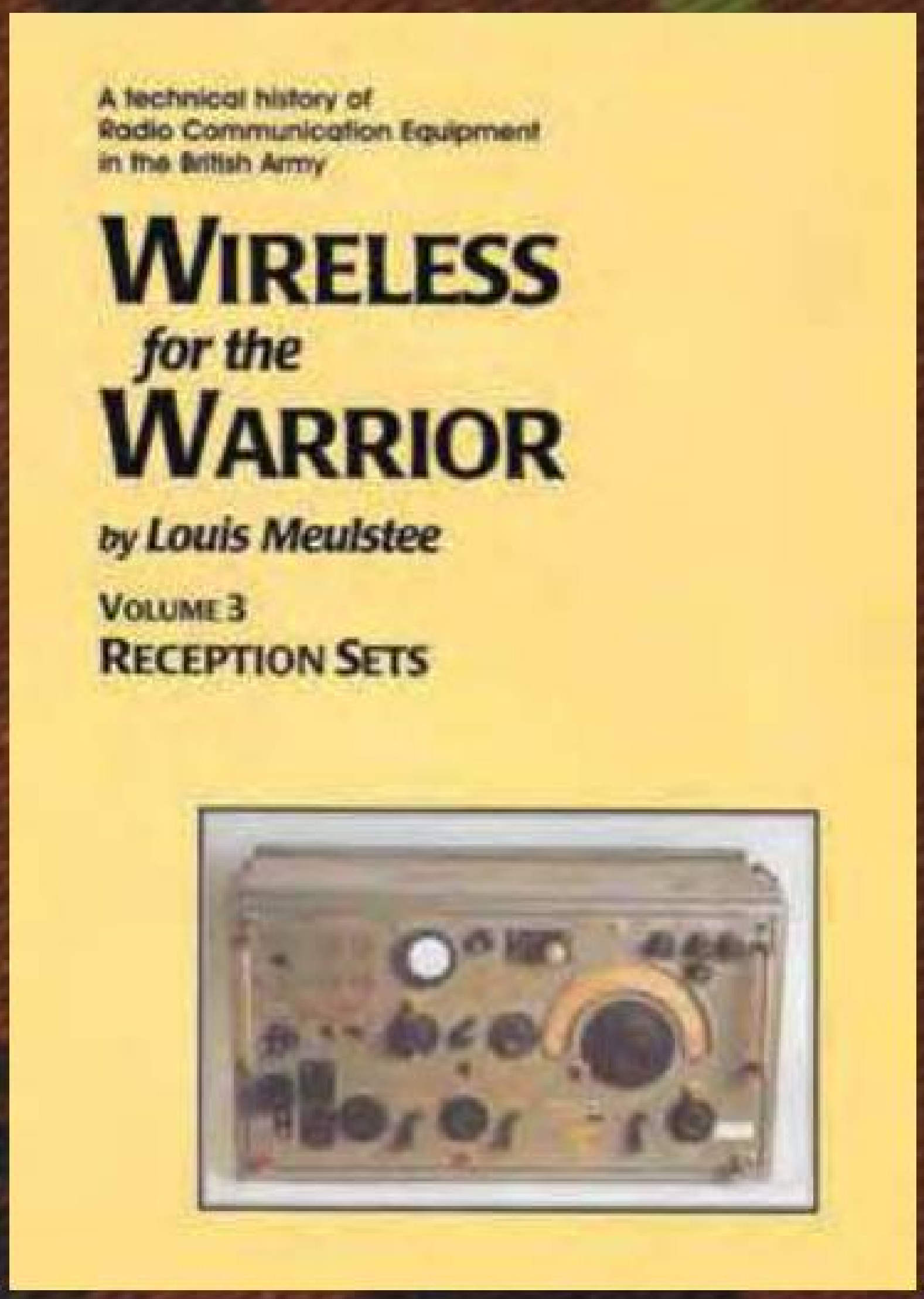

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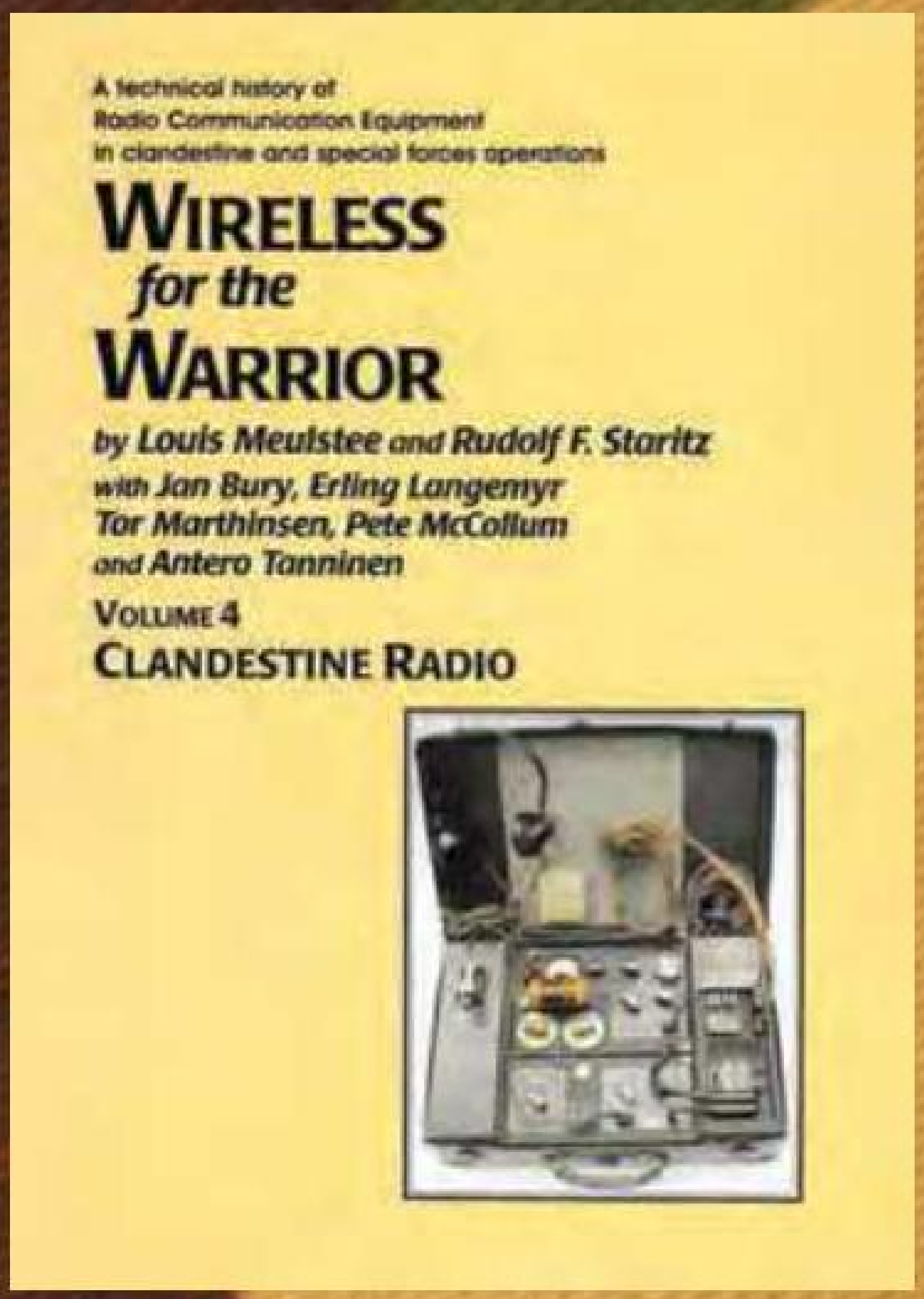

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Editorial

On yer bike 2!

Apologies for recycling a headline from last month, but we seem to be on a bit of a cycling roll here at *PE*. Regular *PE* columnist and recycling guru Julian Edgar has come up with a superb two-part project on converting an exercise bike into a mini-generator. His develops a third of kilowatt, which is pretty impressive. It's just the kind of useful and imaginative project we enjoy at *PE*, and if you build your own version do please send us a photo.

Gemini amplifier

I'm grateful to reader Mike Lister who wrote in after we asked if there were any Gemini amp builders still reading *PE*. Mike writes, 'I went into my loft to get mine out, and it still works! I'm sorry the front panels are so grubby – but they are 50-years old! I still have four of the magazines that the project ran in but seem to have lost the fifth, which I seem to remember had information about ordering all the parts as a kit, which is what I did. It took a long time to arrive as the project was so popular that they ran out of some components, but eventually I was able to construct it and the sound was very good indeed. It drove my two Quad electrostatic speakers perfectly.

'I used the system all the time until I moved house about 15 years ago. I didn't have room for the enormous speakers so settled for a smaller system. The Gemini then got relegated to the loft until your editorial, which prompted me to get it out and have a look. I hooked it up to some speakers and it still sounds good all these years later.'

Thank you for sharing your amp story Mike, and also for offering to help Steve Moreham complete his – I have passed on your details.

Summer Sale

Last, a reminder that our summer sale finishes on 30 September – so do please see the ad on page 11 outlining the details, and then head over to the online store for some PCB bargains!

Keep well everyone

Matt Pulzer

Publisher

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The **Fox** *Report*

Barry Fox's technology column

More lessons from lockdown

One benefit of lockdown has been the enforced gift of time to fix tricky tech problems that had previously been set aside. A Windows PC almost always has some tricky problem to fix – often caused by one of the seemingly endless and hugely time-consuming updates which Microsoft is now pushing out, apparently with far less pre-push testing than previously.

If it ain't broke…

Charmingly, the more major pushed downloads are officially described as 'Feature Updates', with the most obvious 'feature' often being the breaking of something that didn't need fixing.

Sometimes things just stop working – the last update (Ver. 2004) killed USB connection to my Belkin Compact Flash card reader. Luckily, I had another brand reader which still works – annoying, but no big deal.

However, twice over the last year or so I've had working PCs bricked by Windows updates. Most recently, the all-important Start menu stopped working, with a 'critical' error message. As per usual, the message half-promised a self-fix after a task re-start or full re-boot. As per usual, it didn't self-fix; and also, as per usual, a previously made system 'restore points' failed to help.

Googling for community help delivered many wildly different suggested fixes. In fact, the wide variety of possible fixes says it all – the 'Wonderful World of Windows' is now so thoroughly screwed up that no one, including Microsoft, can predict what problems an update will cause until it has been pushed out to the guinea pigs – us.

Restore

The main lesson I have learned the hard way is that it will usually be quicker and easier to restore from a full system backup than get sucked into the time-consuming swamp of trying a string of randomly recommended 'may work' fixes. Just try re-booting the PC, and follow any simple recommendations, then give up and copy all your data off to a separate hard disc drive and do a full system restore.

Strategy for survival

But for this to work, you will need to have adopted a survival strategy while Windows is working 'normally' (ie, with only the usual crop of minor problems); here's mine.

Make a note of all activation keys for Windows and all proprietary software; also, all passwords. You never know when a message will pop up after an update asking for a key or PIN.

Install simple backup software that lets you make regular full system images of the PC to a separate hard drive. Progressively delete older backups to make room for new ones, once a month – or better still, once a week. Always keep at least a couple of the most recent ones.

I always recommend the excellent Macrium Reflect. There is a free version and it can be set to provide an option to restore as the PC boots, and before Windows has a chance to take over and start displaying assorted error messages.

Inevitably, restoring from a week-old, or month-old, backup will leave you with a system that is a week or month out of date. This is why you need to copy all data to a separate HDD before starting the system restore.

Even if the Start Menu has vanished, you can still copy data to a separate HDD if you have placed a short cut to 'My Computer' on the desktop. This lets you navigate round the PC's storage zones without a Start Menu. I found this purely by chance. If I had not previously put a My Computer shortcut on my desktop, for who knows what reason, I'd probably have been stuffed.

Doubtless some readers will have their own ways of automating regular data back-up. Fine, but always

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remember that you never know for sure whether a backup is good until you need it. Data copied manually to a separate drive is easy to check, by clicking and opening a few files.

A new strategy

Recognising that Windows will continue to make my life difficult; I have now implemented a new routine. All my data is now stored on a separate hard drive, with shortcuts to the drive and stored files and folders put on the desktop. So, doing a full system restore from a week or month ago no longer involves rebuilding data.

Put simply I have now banished my Windows operating system to its own patch, without any of my everyday data.

Go open source

I am also adopting a general policy of using only open-source key-free software wherever possible. Fortunately, more and more open-source alternatives to proprietary key-dependent programs and apps are becoming available. The proprietary developers who lose sales should blame Microsoft for making Windows so prone to update problems.

As I write, I can hear Mac users chortling. Mock ye may! Recent Mac updates have knocked out 32-bit software – without a pre-install check and red flag warning, and without an easy option to roll back to its previous 32-bit-friendly state. Has Apple been taking lessons from Microsoft in how to lose friends and alienate customers?

Update on A-to-D content rescue

I recently heard from reader Ralph Bacon, who wrote: ‘I was very interested in your article in September 2020’s *PE* regarding A-to-D content rescue. Not for the actual analogue-video-tapes-to-digital rescuing (although I have many such tapes) but the final paragraph made me sit up.

‘About four years ago I bought a Magewell HDMI-to-USB converter, currently on sale at Amazon for £289.50 (see: **https://amzn.to/2QkVhcV**). When I bought it, it was a snip at £250! It’s worked fine for me every week since I bought it – thank goodness. Nevertheless, I searched for the low-cost HDMI-to-USB converter item you mentioned, thinking it must be a scam – but no!

‘Less than seven days ago I bought an (over-named) ‘USB 2.0 3.0 4K Loop Out Audio Video Capture Card HDMI Recording Box Mic In Phone Game Live Streaming for Switch PS4 DVD Camera’ on the YiBao Trading Ltd Store at AliExpress for the princely sum of $10.38, shipping included – see: **http://bit.ly/pe-oct20-ali**

‘It arrived today (six days later) and it really is a USB 3.0 HDMI capture device and seems to capture my video just as well as the Magewell. It needs a bit of longer-term testing, but how can they do this? £289 versus $10 – what magic have they discovered? A new chip, perhaps? If you get yours before Christmas I’ll be interested in what you discover as I have no intention of tearing mine down!

‘So, thank you very much for the heads-up. It now means that an expensive, single-point-of-failure in my video setup has just been eliminated. Fantastic!’

And thank you Ralph for the tip – mine has arrived and I also find it too good to be possible! I am still testing, and will write more later, but one early word of caution. It does not have the hardware processing power built into more expensive devices, you need a powerful PC to drive it. Also, it only handles PCM audio.

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HP8341A	Synthesised Sweep Generator 10MHz – 20GHz	£2,000
HP83630A	Synthesised Sweeper 10MHz – 26.5 GHz	POA
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Marconi 2955	Radio Communications Test Set	£595
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Racal 1991	Counter/Timer 160MHz 9 Digit	£150
Racal 2101	Counter 20GHz LED	£295
Racal 9300	True RMS Millivoltmeter 5Hz – 20MHz etc	£45
Racal 9300B	As 9300	£75
Solartron 7150/PLUS	6½ Digit DMM True RMS IEEE	£65/£75
Solatron 1253	Gain Phase Analyser 1mHz – 20kHz	£600
Solartron SI 1255	HF Frequency Response Analyser	POA
Tasakago TM035-2	PSU 0-35V 0-2A 2 Meters	£30
Thurlby PL320QMD	PSU 0-30V 0-2A Twice	£160 – £200
Thurlby TG210	Function Generator 0.002-2MHz TTL etc Kenwood Badged	£65

HP/Agilent HP 34401A Digital
Multimeter 6½ Digit £325 – £375

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HP53131A	Universal Counter 3GHz Boxed unused	£600
HP53131A	Universal Counter 225MHz	£350
Audio Precision	SYS2712 Audio Analyser – in original box	POA
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Time 9818	DC Current & Voltage Calibrator	POA

Marconi 2955B Radio
Communications Test Set – £800

Now is the autumn of our discontent...

Techno Talk

Mark Nelson

…*not* made glorious by this arsonist of Lancashire (apologies to The Bard), who is now serving three years behind bars at Her Majesty's pleasure for torching a Vodafone base station in Knowsley. His excuse was that he wanted to 'put the mast out of action' because 5G technology helps transmit coronavirus.

How silly can you get? If he had only bothered to check the coverage page on Vodafone's website he would have seen this out-of-town base station was not equipped for 5G, which Vodafone has so far only rolled out 'in select areas of large towns and cities'. An analysis of his phone revealed he had instead wasted his time in online chat groups about 5G technology.

Myth-understandings

As you doubtless know, theories claiming that 5G technology helps transmit coronavirus have been condemned widely by the scientific community. Indeed, they have been branded 'the worst kind of fake news' by NHS England medical director Stephen Powis. According to the BBC, the theories fall broadly in to two camps: either that 5G can suppress the immune system, making people more susceptible to catching the virus, or else the virus can somehow be transmitted by means of 5G technology. Both of these notions were condemned as 'complete rubbish' by Dr Simon Clarke, associate professor in cellular microbiology at the University of Reading.

You could not make this up

In reality, the connection is far more subtle and consequently much less obvious, as someone revealed to me exclusively for this article. This is the real deal, in his words, *not* mine. 'If you have been Covid-19 checked, you have been chipped. The swab is pushed to the back of your nasal passages so you cannot see it. You may notice a very tiny black dot on the swab going in and not when it comes out. That is the microchip, which runs on the human electric current just like you see on an electrocardiogram. Once implanted, you have to use a magnifying TV camera to push inside to find it and by using the very long swab they can implant the chip without you knowing it. The chip, equipped with an ultra-thin antenna wire that you cannot feel, makes it possible to track you everywhere because it does not require a satellite but instead runs on 5G cell phone networks for undetected communications. It relays any conversation it can pick up with a sensitive microphone. A symptom is you may begin sneezing violently as your sinuses try to get rid of it, hence the wound to hide it. Well, that's what they told me when I asked why my cell phone shows a signal all the time, even when I know there is no cell service near me. Welcome to the New World Order, fellow slaves.'

Of course, not a word of the above is true, but it's exactly the kind of highly inventive and almost-plausible fantasy that is spouted on the more conspiracy-obsessed corners of the Web.

The things you find on the Web

Everything you want to know is on the Internet. The trouble is that most of it is not indexed, such as this nugget. If you work in a radiology department at a hospital, don't bring your iPhone with you. I quote: 'Apparently iPhones were failing when used near MRI scanner areas. The problem was not the magnets themselves, but the outgassing of large quantities of helium gas when the units were serviced or the magnet quenched. The iPhones have a MEMS oscillator timing device inside and the helium was migrating into the MEMS package past the seal and weakening the vacuum inside. This lowered the operating frequency.' True? Who knows?… and that's the problem.

And finally…

Let's end on a positive note, with a fascinating example of how a perceived limitation may actually not be the constraint you imagined. Take the capacity of a smartphone battery, typically somewhere between 1500 and 2000mAh. If that's what it is, surely the laws of physic say you're not going to get any extra performance than what it says on the label.

But what if you're a contrarian and decide to think outside of the box? That's precisely what researchers at the University of Essex did in order to squeeze extra life from their batteries. Their new research methodology employs machine-learning algorithms to optimise a mobile phone's performance and thermal behaviour based on the user's interaction with the phone. In lab tests, this entirely novel technique has outperformed existing methodologies in the top smartphones used for the purpose of performance and thermal behaviour.

Developed by a team led by computer scientist Somdip Dey, the technique would enable users to use their phone for longer and mitigate the ongoing issues with smartphone battery life. Their work has already gained interest from other researchers keen to pursue the methodology and further the development of resource-optimisation techniques in mobile phones.

'This is ground-breaking work and the first to propose reinforcement-learning based on a machine-learning approach to optimise performance, energy consumption, and thermal behaviour in a mobile device by taking the user's behaviour with the device into consideration,' explained Somdip, who pursued the work while working at the Samsung R&D Institute last year. 'By learning from the user's behaviour, we have shown how smartphones could be developed to get even more battery life for the same usage.'

At lunchtime a user might quickly scroll through the BBC News app checking the headlines, which will require a higher FPS (frames per second) than when they spend more time on the app in the evening, slowly scrolling down and reading more stories in full. The researchers' methodology detects the change in FPS for the app being used and tries to find the best operating frequency of CPU and GPU processors to cater for the change in app-use behaviour, while aiming to consume the least amount of power and minimise the temperature rise in the device, which is a critical issue in mobile phones.

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Net Work

Alan Winstanley

This month, *Net Work* ranges from a spookily accurate prediction of tech use from the 1940s, to the latest plans for visiting the Red Planet and the launch of microsatellites.

A fascinating video recently surfaced on YouTube entitled *La Télévision, œil de demain* (*Television, eye of tomorrow*) by JK Raymond Millet (1947). The prophetic French film shows ordinary people walking around and travelling on public transport while viewing a smartphone-like TV screen. Engrossed by these devices, they bump into each other or walk in front of traffic. Sound familiar? You can view it on: **https://youtu.be/ZKfOcR7Qbu4**

Some 35 years later, we reported in our November 1982 issue that the world's longest optical fibre telephone cable had just come into service in Britain. Located between London and Birmingham, and measuring 205km (128 miles), it was the product of British Telecom (BT) scientists who developed the precision fibre manufacturing process critical to transmitting data efficiently over long distances. We foretold that tomorrow's communication 'highways' in Britain, carrying thousands of phone calls, computer data and TV pictures between towns, would consist of hair-thin strands of glass like these. BT's 'Lightlines' technology would go on to be licensed to British, European and US companies and 'fibre' is now the byword for high-speed Internet access. It is may be no surprise that a besieged Huawei intends to base its new international fibre optic research HQ right here in the UK.

Google has just announced the construction of its fourth privately funded undersea fibre optic cable to carry data between the US, the UK and Spain. Due for completion in 2022, the British end of the 6,300km cable, named 'Grace Hopper' to commemorate the work of computer scientist Grace Brewster Murray Hopper, will terminate in Bude in the south-west county of Cornwall – the same county where in 1901 Marconi used a 13kW spark transmitter to send the world's first transatlantic radio signal (the letter 's' in Morse) to Newfoundland. Details of the Marconi Centre and some history are at: **http://bit.ly/pe-oct20-marconi**

Fibre optics play essential roles in ground-based telecoms networks and the gradual roll-out of 'fibre to the property' (FTTP) will enable large volumes of 'hyperfast' Internet data to be delivered to the home and workplace. Faster Wi-Fi networks are also in the pipeline (see later).

In data centres around the world, capacity has spiralled upwards as well, allowing 'cloud computing' to manage everyday online services. For most of us, cloud computing means that complex data processing, web content management, social media and services such as search or email are hosted off-site, but we're not sure where, nor do many of us worry about it. The data centres themselves might be in Washington, Tyne & Wear or Washington DC for all we know, and of course they are heavily interconnected by satellite and fibre optics to spread the load and maintain uptime.

Held to ransom

As Internet users, we casually upload often-sensitive data onto the web and we trust the service providers to take good care of it, wherever it may be hosted. In May this year, the cloud-computing service provider Blackbaud was the subject of a ransomware attack in which a subset of user data was exfiltrated by blackmailers. Blackbaud hosts online services for countless 'good causes' like universities, charities, fundraisers and political groups around the world, but news of the ransomware attack only broke in the UK several months later when, in July, it was learned that wide-ranging student personal data hosted on behalf of mainstream UK universities had been stolen. The breach also hit the UK's National Trust, which safeguards Britain's historical monuments, parks and gardens; US and Canadian institutions were affected as well. The impact of the data breach is still emerging several months later, but the UK Information Commissioner's Office (ICO) has so far been silent about the incident.

Honour among thieves?

Blackbaud PR claimed that it had detected the intrusion and intervened to prevent large-scale file encryption on its servers. Extortionists then demanded a ransom to supply the decryption tool (maybe). Cybercriminals may even leak some stolen data onto a website in order to tighten the thumbscrews. Authorities including the FBI and police remind us that paying a ransom is never recommended: it merely encourages more cybercrime and there is never any guarantee that you will get your data back anyway. One cannot

Glued to his screen: a prescient view of a French future mobile technology user in *La Télévision, œil de demain* (1947) – JK Raymond Millet. (Image: YouTube)

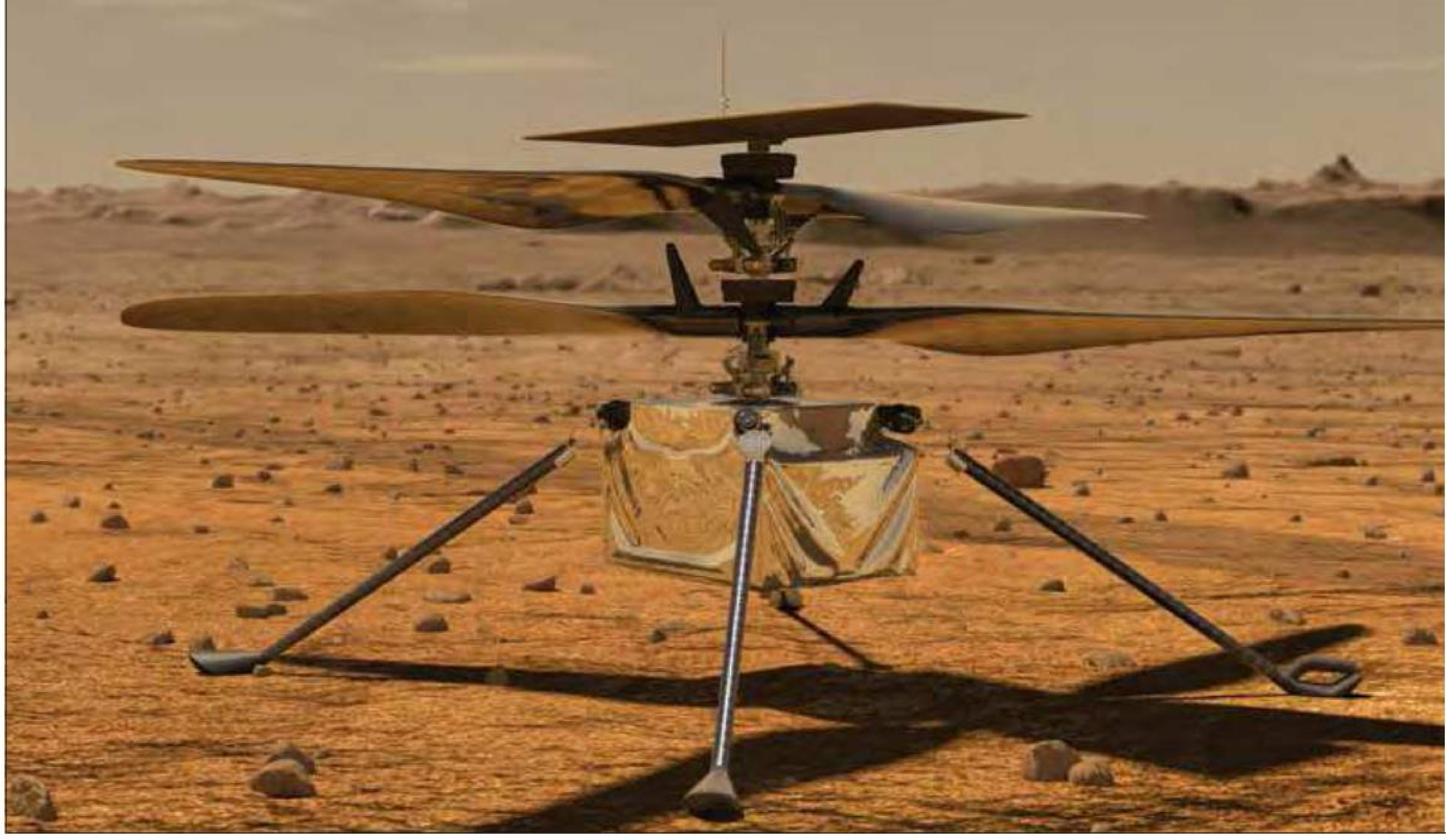
Google's Grace Hopper is its fourth undersea fibre optic cable and will connect the US, UK and Spain (Image: Google) know whether stolen personal data is likely to end up for sale somewhere on the dark web.

However, the cybercriminal marketplace has moved on in the past year or two, and realists recognise that payment of a ransom is now deemed an option of last resort. Blackbaud did indeed pay up. 'Because protecting our customers' data is our top priority, we paid the cybercriminal's demand with confirmation that the copy they removed had been destroyed. Based on the nature of the incident, our research, and third-party (including law enforcement) investigation, we have no reason to believe that any data went beyond the cybercriminal, was or will be misused; or will be disseminated or otherwise made available publicly,' said a Blackbaud PR statement.

They have been far from alone in suffering fallout from a ransomware attack. In late July, navigation gadget maker Garmin was also hit by malware which crippled their online platforms. The WastedLocker file encryptor is considered impregnable and the only way to unlock files is to buy the corresponding decryption key, according to tech website *Bleeping Computer* which simulated the Garmin attack and decryption using sample files before concluding that a ransom must have been paid for Garmin services to be restored so quickly afterwards (see their expert analysis at **https://tinyurl.com/y59kjwk7**).

Ransomware can cripple small businesses, as I have witnessed first-hand, but larger enterprises are insured against such losses and a new sector of threat management has evolved. Specialist firms such as Arete Incident Response (**https://areteir.com**) are employed to monitor and investigate a data breach and will even negotiate a Bitcoin settlement directly on their client's behalf. Perversely, it's possible that cybercriminals, seeing the millions of dollars that can be made from their racket, are on a reputation management drive of their own, as they reposition themselves as 'honourable crooks' keen to do business with victims and offering them a reliable service in return for a ransom.

In my book, ransomware remains the Number 1 threat to Internet users because of the serious damage it can do when valuable user data is stolen or goes up in smoke, whether it's hosted on your PC or somewhere in the cloud. Regular readers will recall my advice to remain vigilant and take a 'last gasp' data backup onto a pocket drive or SSD. Even data backed up onto a DVD is better than nothing. If a computer drive suffers an attack, it's probably best replaced with a new one rebuilt from scratch.



An artist's concept shows the Mars *Ingenuity* helicopter on the Martian surface. (Image: NASA/JPL-Caltech)

The TikTok countdown

Huawei continues to feel the force of brutal US sanctions to choke off component supplies to the privately owned Chinese tech giant, citing risks to national security. Huawei's smartphone business in the West has undoubtedly been hit by the loss of Google services. However, emerging from the Coronavirus pandemic, Chinese industry has sprung back to life and Huawei's global smartphone sales surpassed Samsung for the first time. 'It marks the first quarter in nine years that a company other than Samsung or Apple has led the market, a remarkable result that few people would have predicted a year ago,' said Canalys Senior Analyst Ben Stanton. The boost is attributed to higher sales in China, up 8%. 'If it wasn't for COVID-19, it wouldn't have happened.' That picture may change once the global economy ramps up to speed again.

Meanwhile, the US has turned its attention to TikTok, the popular Chinese-owned platform that hosts short video clips. TikTok and scores of other Chinese apps have already been banned in India for 'security reasons' with the risk, consistently denied by TikTok, that sensitive user data is sent to China. Geo-politics is probably involved after the two countries skirmished on a disputed border. The US has since moved to shut down TikTok for similar security reasons but has granted TikTok 45 days to work out a takeover deal with Microsoft for some operations that could be worth up to $50bn. As if the dynamics aren't complicated enough, TikTok is also investigating siting its new HQ in London.

Ingenuity Unlimited

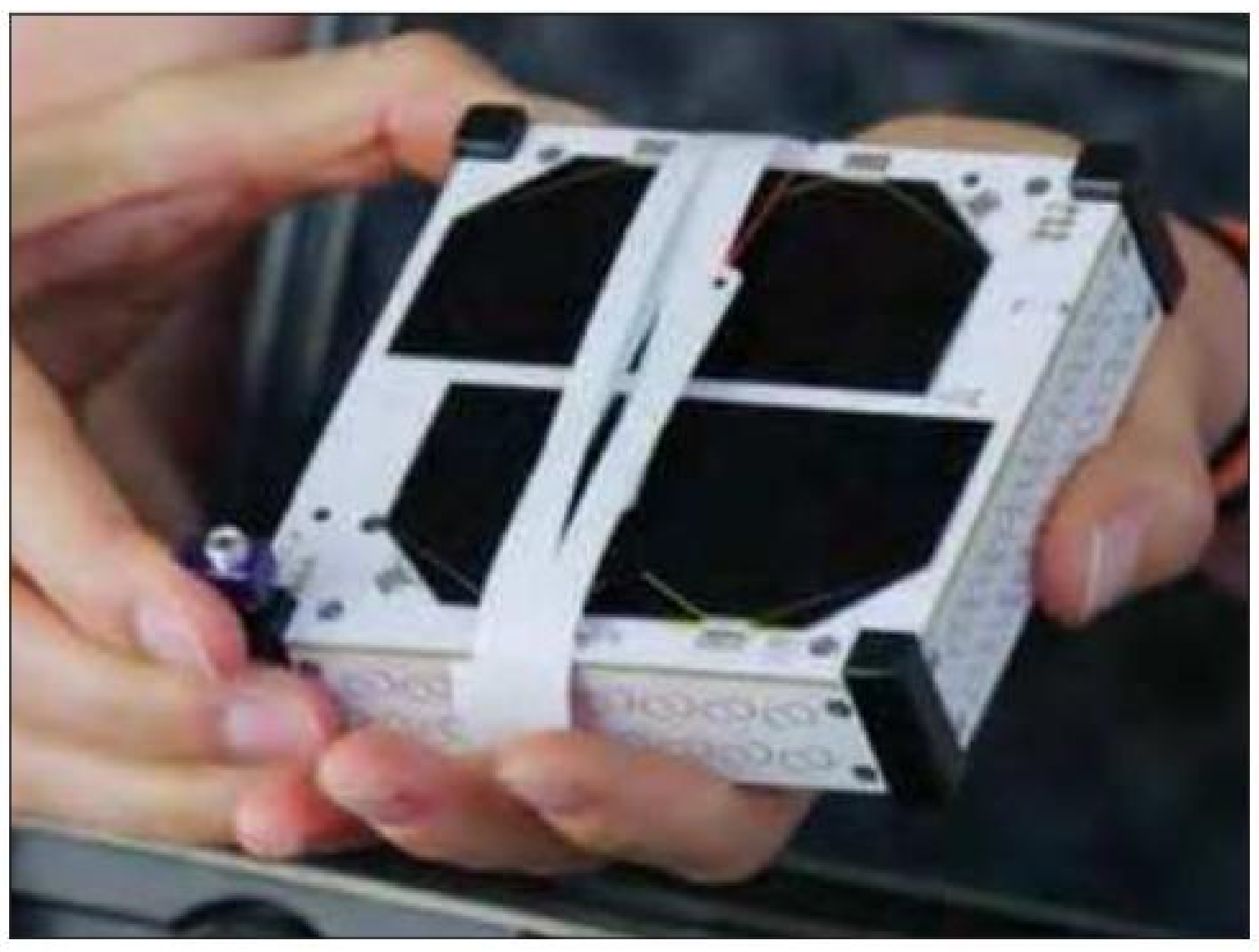
In July, NASA successfully launched its latest Mars-bound explorer called *Perseverance* in their Mars 2020 program (see: **https://youtu.be/JIB3JbIIbPU**). The Martian rover is the largest and most sophisticated vehicle NASA has ever built, and it carries a proof-of-concept helicopter called *Ingenuity* that will test-fly through the Martian atmosphere. *Perseverance* is destined to land in February 2021.

Back in July 2019's *Net Work* I mentioned how Amazon was recruiting for its Project Kuiper, which aimed to, 'offer high-speed broadband connectivity to unserved and underserved communities around the world' using a constellation of no less than 3,200 satellites. It has now received FCC approval to proceed and eventually its low-earth orbit (LEO) satellites will compete for business with SpaceX's Starlink service. SpaceX also celebrated the first commercial launch and safe return of two astronauts who spent 63 days on board the ISS (International Space Station). The return flight was the first US splashdown in 45 years. Four astronauts are scheduled for the next manned mission in October.

How small can a satellite be? Swarm Technologies (**www.swarm.space**) is a relatively new satellite company offering lowest-cost global connectivity through its own pint-sized 'SpaceBEE' satellites. They claim to operate the world's smallest satellites with target markets of Internet of Things, maritime, 'connected cars' and agricultural data transmission in mind. Just as rack-mounted appliances like network servers are based on standard '1U' sizes, so-called CubeSat nanosatellites really are tiny – here, 1U is just 10 × 10 × 10cm (see **https://go.nasa.gov/3hpZ5p5**). Swarm's SpaceBEEs measure even less, a diminutive ¼U and fit in the palm of your hand. Swarm expects to launch 150 SpaceBEE microsatellites and build up a network of 30 ground stations, with SpaceBEEs hitching a rideshare on any rocket platform as they become available. The microsatellites last up to two years.

Other news

Ofcom is making the lower 6GHz band available for next generation Wi-Fi and



Swarm's SpaceBEE is a tiny '¼U' microsatellite that will handle data for IoT, agriculture, marine and similar dedicated tasks. (Image: Swarm Technologies)

Radio LAN (RLAN) use, on a licence-exempt basis. They hope to increase speeds and reduce Wi-Fi congestion locally and promote the development of new technologies. 'Wi-Fi 6' has been compared to advances in 5G mobile telecoms, and network devices built to the 802.11ax standard are expected to follow in due course. Upgraders will therefore have to check their specs for 2.4GHz, 5GHz and 6GHz Wi-Fi. Currently, dual-band routers offer 2.4Ghz and 5GHz frequencies, but a so-called 'triband' router offers one 2.4GHz and two 5GHz radios, which may be overkill for average Internet users. A great summary of 802.11 tech specs can be found at: **http://bit.ly/pe-oct20-wifi**

If you own a Netgear router or extender, then it's worth checking to see if it's on Netgear's list of devices that have reached end-of-life and will not receive any further patches or security updates. They are classed as 'outside security support period' and the risk from unpatched vulnerabilities means that it may be time to source a more modern replacement; see: **http://bit.ly/pe-oct20-netg**

BT has launched a new broadband service which provides busy households with a second broadband service, enabling residential customers to split work or schooling (say) and gaming or video-conferencing traffic into two network connections. BT's 'Dedicated Connection' includes a Smart Hub and a home visit from a BT tech expert and costs £49.99 a month or £59.99 for non-BT customers. On 1 September most existing BT phone customers are also being pushed onto a new 'Unlimited Minutes' calling plan offering unrestricted calls to UK landline and mobile numbers for a fixed price of £4.50 a month, though other pricing options will still be available.

Following my item last month on buying IP cameras, another option worth checking is whether cameras offer ONVIF compatibility. ONVIF (Open Network Video Interface Forum) is a non-profit industry group aiming to offer interoperable security standards. Independent ONVIF-compatible apps and software can be viable alternatives to paid-for IP camera software licences such as Synology Surveillance Station. A free ONVIF viewer app allowed the author to dedicate an older Android tablet to checking a Ucam IP Camera (see last month) when the maker's app wasn't suitable.

Twitter was recently the subject of a highly embarrassing attack when social engineering techniques were used over the phone on some key Twitter employees that ultimately led to 130 prominent users' accounts being hacked, including those of Bill Gates, Barack Obama and Elon Musk. Bogus philanthropic tweets were then sent to solicit $1,000 Bitcoin payments in return for supposedly receiving back double that amount. In the space of just two weeks, three hackers were traced and have since been charged, including a 19-year old British man.

See you next month for another *Net Work* roundup!

The author can be reached at:
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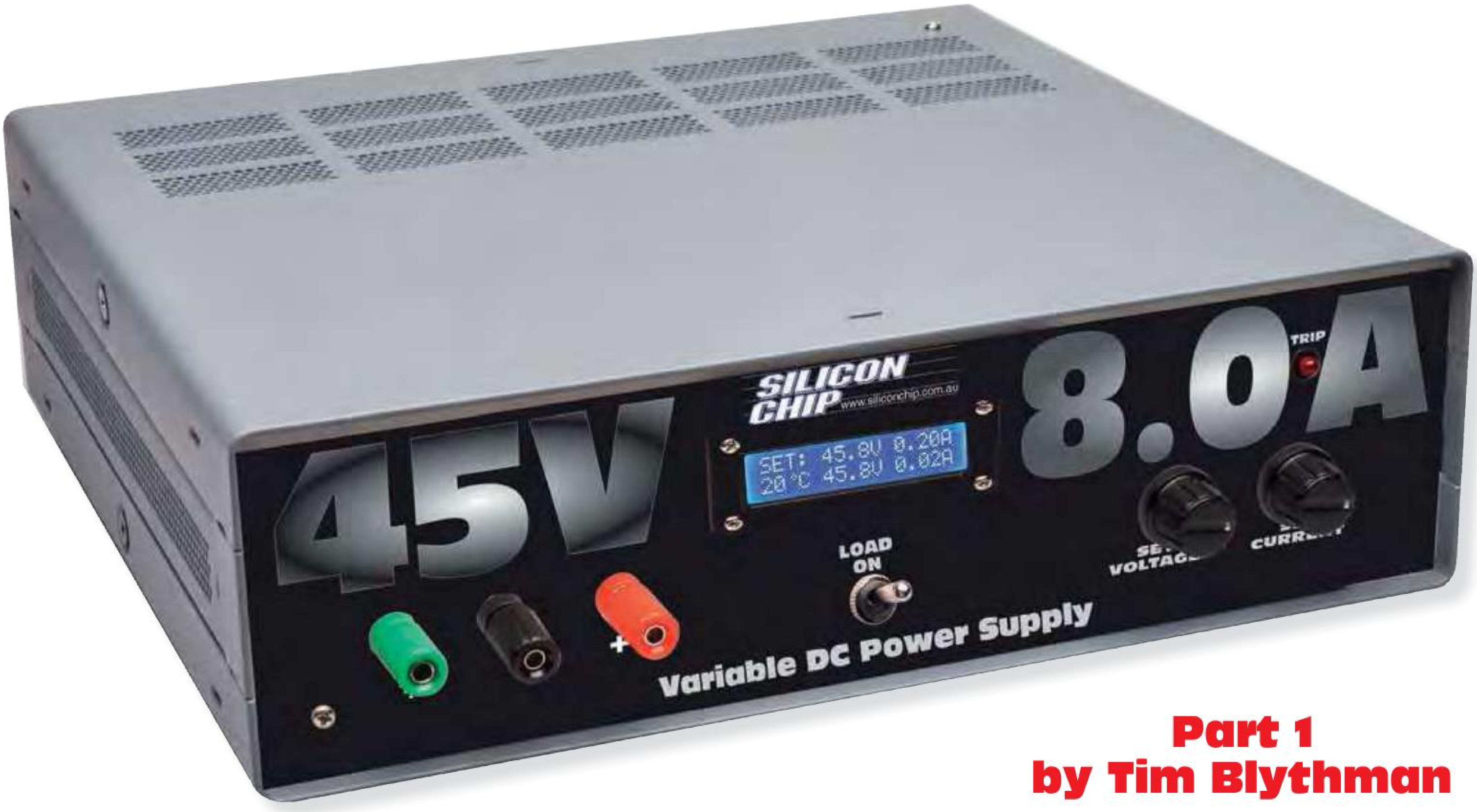
HIGH-POWER 45V/8A VARIABLE LINEAR SUPPLY



Part 1

by Tim Blythman

This adjustable bench supply can deliver heaps of power, up to 360W in total, making it ideal for the test bench or just general purpose use. It can operate as a voltage or current source at 0-45V and 0-8A. It is an entirely linear, analogue design. It's fan-cooled with automatic fan speed control, short circuit/overload protection and thermal self-protection. It can even be used as a basic but powerful battery charger.

We've been designing this project for quite a while. However, it has taken some time to get it just right – but now, it's finally here.

This power supply can deliver up to 45V at up to 8A, or up to 50V at lower currents. It has a fully adjustable output voltage down to 0V and an adjustable current limit. Its operating envelope is shown in Fig.1.

That makes it suitable for many different tasks, including testing newly built or repaired equipment, temporarily running various devices and even charging batteries.

Its controls are simple. Two knobs set the voltage and current limits, and the power supply maintains its outputs within these constraints.

It displays on an LCD screen the *Supply's* actual output voltage, set voltage, actual current, set current and the heatsink temperature.

These can be shown on an alphanumeric LCD, or if you prefer, you could use separate LED or LCD panel meters.

It has a pair of internal high-speed fans to keep it cool. These automatically spin up and down as required. If the *Supply* is operated in the orange shaded area shown in Fig.1, or at very high ambient temperatures, or the fans fail, a thermal current limit comes into play. This reduces the output current until the unit cools down, preventing damage to the *Supply*.

While we had originally planned for this power supply to be able to deliver 50V at 8A, it is difficult to achieve that with a practically sized transformer and a reasonable parts budget.

It's limited to 45V at 8A because despite using a large 500VA transformer, its output voltage still sags significantly under load, meaning there isn't enough headroom for regulation.

However, if the transformer was upgraded (and possibly the filter capacitors too), then it could be capable of delivering the original target output of 50V at 8A.

Features and specifications

- Up to 45V output at 8A, 50V output at 2A (see Fig.1)
- Low ripple and noise
- Adjustable output voltage, 0-50V
- Adjustable output current, 0-8A
- Constant voltage/constant current (automatic switching)
- Shows set voltage/current, actual voltage/current, heatsink temperature
- Fan cooling with automatic fan speed control
- Thermal shutdown
- Fits into a readily available vented metal instrument case
- Switched and fused IEC mains input socket
- Uses mostly commonly available through-hole components

Design overview

The basic design of the *Linear Supply* is shown in the simplified circuit diagram, Fig.2. It's based around an LM317HV high-voltage adjustable regulator, REG3. The LM317HV variant can handle up to 60V between its input and output, at up to 1.5A.

Clearly then, this regulator cannot pass the full 8A output current. And even if it could, it couldn't dissipate the 400W that would be required (50V × 8A) as it's in a TO-220 package.

Therefore, the regulator itself only handles about 10mA of the load current, with the rest being delivered by four high-power current-boosting transistors, Q4-Q7.

Power is fed into the supply via the IEC input socket shown at upper left, and passes through the mains switch and fuse before reaching the primary of transformer T1.

Its two 40V AC secondary windings are connected in parallel and then on to bridge rectifier BR1 and a filter capacitor bank, generating the nominally 57V DC main supply rail.

This passes to the input of REG3 via a resistor, and also to the collectors of the NPN current-boosting transistors and the emitter of PNP control transistor Q3.

As the current supplied by REG3 rises, Q3's base-emitter junction becomes forward-biased, and it supplies current to the bases of Q4-Q7, switching them on.

As REG3 draws more current, they switch on harder, providing more and more current to the output of the supply. These transistors therefore supply virtually all of the maximum 8A output current.

Regulator control

Like most adjustable regulators, REG3 operates by attempting to maintain a fixed voltage between its output (OUT) and adjust (ADJ) pins. In this case, around 1.2V.

Usually, a resistor is connected between OUT and ADJ, and another resistor between ADJ and GND, forming a divider.

As the same current flows through both resistors, the voltage between ADJ and GND is fixed, the regulator output voltage is that voltage plus the 1.2V between OUT and ADJ.

But in this case, rather than having a fixed or variable resistor from ADJ to GND, we have transistors Q1 and Q2, connected in parallel. Their bases are driven from the outputs of op amps IC1a and IC1b. Their emitters go to –5V so that the ADJ pin can be pulled below ground, allowing the regulator OUT pin to reach 0V.

This is important both to allow low output voltages and for the current limiting to be effective.

Op amp IC1a compares the voltage from the wiper of the VOLTAGE SET potentiometer to a divided-down version of the output voltage. It provides negative feedback so that if the output voltage is higher than the setpoint, Q1 is driven harder, pulling the ADJ pin of REG3 down, reducing the output voltage.

If the output voltage is too low, Q1's base drive is reduced, allowing REG3 to pull the output up.

A capacitor from the ADJ pin of REG3 to the –5V rail helps to stabilise this arrangement. Current control op amp IC1b and its associated transistor Q2 work similarly, to regulate current. Because transistors Q1 and Q2 can only sink current, the output voltage will be determined by which is lower: the voltage setting, or the voltage required to achieve the desired current setting.

The output current is monitored via a 15mΩ (milliohm) shunt between the output of REG3/Q4-Q7 and the output terminal. Voltage feedback comes from the output side of this resistor, so the supply will automatically compensate for the shunt's voltage drop (up to 120mV).

Shunt monitor IC4, a form of differential/instrumentation amplifier, converts the voltage across the shunt to a ground-referred voltage so that IC1b can compare it to the voltage from the current set pot.

By using control voltages to set the desired output voltage and current, we can easily show these on the front panel of the meter, so you can see what you're doing.

LM317-type regulators have a minimum output load current, which is provided by a constant current sink comprising transistors Q8 and Q9.

Otherwise, the output of REG3 would rise of its own accord. The current sink dissipates a lot less power than a fixed resistor would, as the resistor would draw much more current at high output voltages.

The NTC thermistor on the heatsink forms a divider with a resistor such that the voltage at their junction drops as the temperature increases.

This voltage is fed to a PWM generator which increases the duty cycle fed to the gate of MOSFET Q10 as the temperature increases, speeding up the two 24V fans.

The fans are connected in series and run from the 57V supply rail via a dropper resistor. This is a much more power-efficient arrangement than running the fans from one of the regulated rails.

The temperature signal is also fed to control logic which biases NPN transistor Q12 on if the heatsink gets too hot, pulling the current control signal to ground and shutting down the supply.

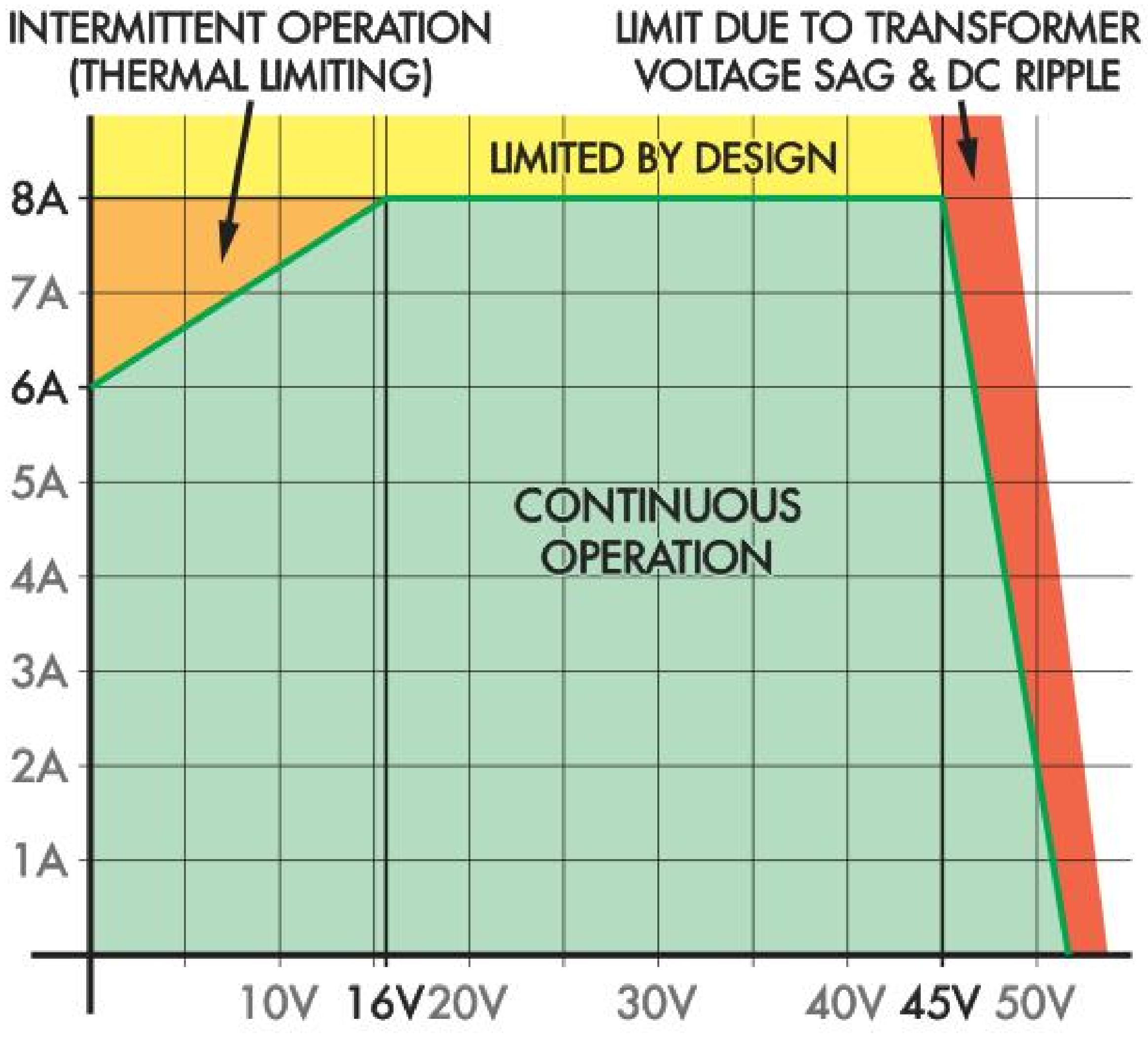


Fig.1: the *Linear Supply* can deliver 8A but can only do so continuously with an output voltage of between 16V and 45V. Below 16V, internal dissipation is so high that the unit will go into thermal limiting after a few minutes. Above 45V, transformer regulation means that the DC supply voltage drops far enough that 100Hz ripple starts appearing at the output, so the actual voltage may be lower than the set voltage.

Here's a teaser look inside our new *Linear Supply*, taken before we applied the dress panel. Full construction details will begin next month. As you might expect from its specifications, there's a lot to this supply, dominated by the 500VA transformer at left. But the good news is that it uses mostly through-hole components so construction isn't too difficult.

Several internal regulators are shown in Fig.2, at upper right. These are required to generate various internal control voltages and to power the control circuitry itself.

The output of the +12V regulator is fed to a capacitor charge pump (IC3) which generates a roughly –9V rail that is then regulated to –5V.

As mentioned earlier, this is needed to allow the *Supply* output to go down to 0V.

Thermal considerations

One of the biggest challenges when designing this supply was keeping it cool without needing a huge heatsink in a massive case. The worst case is when the output is short-circuited at 8A (or it's delivering a very low output voltage at 8A). The required dissipation is then over 400W, and it should ideally handle this continuously.

Three things became apparent during testing:

1) *The current-boosting transistors needed to be mounted on the heatsink with as little thermal resistance as possible, to keep the devices themselves at a reasonable temperature when dissipating around 100W each.*

2) *To keep the heatsink and case size reasonable, powerful cooling fans are required. These should be thermally throttled to keep noise under control.*

3) *The case needs to be vented, with careful attention paid to the airflow paths. We also determined that the current-boosting control transistor (Q3) would need to dissipate over 1W so it too would need to be mounted on the heatsink, along with REG3 and the bridge rectifier, which also dissipates a significant amount of heat at full power.*

Since the heatsink is connected to the collectors of Q4-Q7, which are sitting at 57V, it needs to be isolated from the earthed case, so we came up with a mounting arrangement that achieved this, while still keeping the heavy heatsink nicely anchored.

The fans are sandwiched between the rear of the case and the heatsink, so they draw air through large holes in the rear panel and blow it over the heatsink fins. That air then turns 90° and exits via the pre-punched vent holes in the top/bottom of the case.

This does an excellent job of getting all that heat out of the relatively small enclosure.

Circuit details

The full circuit of the *Linear Supply* is shown in Fig.3. While it is considerably more complicated than the simplified diagram (Fig.2), you should still be able to see how the various sections correspond.

Starting where power enters the input, the 230V AC mains from the input socket/switch/fuse assembly is applied to the two 115V primary windings of 500VA transformer T1, which are connected in series.

The 40V AC from its paralleled secondaries goes to BR1, a 35A bridge rectifier, and from there, to a bank of four 4700μF 63V electrolytic capacitors to carry the circuit over the troughs of the mains cycle.

With no load, the main DC bus capacitors sit at around 57V. The diode drop across the bridge is offset by the transformer's no-load voltage being slightly above nominal. In any case, it is just below the 60V limit of the LM317HV regulator (REG3).

Control circuitry

As mentioned earlier, the LM317HV adjustable voltage regulator (REG3) is the core of the circuit. It maintains the output voltage steady in spite of changes in load impedance and current draw, as long as its ADJ pin voltage is held constant. The ADJ pin is pulled up by an internal current from the input. To regulate the output, the circuit sinks a variable current from the ADJ pin.

This control is exerted by IC1, a dual op amp which runs from a 29V supply, between the +24V and –5V rails. The negative voltage is necessary because the LM317HV's ADJ pin needs to be around 1.2V below the output to regulate correctly. To achieve 0V at the output means that the ADJ pin needs to be around –1.2V relative to GND.

The voltage and current control sections of the circuit around IC1 are quite similar. The reference voltage from the potentiometers is fed into their respective op amp inverting inputs (pins 2 and 6) via 10kΩ resistors, while feedback voltages from the output are fed into the non-inverting inputs (pins 3 and 5) via another pair of 10kΩ resistors.

The user controls the *Linear Supply* via voltage set potentiometer VR3 and current set potentiometer VR4. One end of each is connected to ground so that when set to their minimums, their wipers are at 0V, which corresponds to zero voltage and current at the output.

These are set up as voltage dividers, and both have series 10kΩ trimpots (VR1 and VR2) connected as variable resistors on their high side. This allows you to adjust their full-scale ranges.

The current-setting pot also has a 27kΩ resistor in its divider chain, as the voltage and current adjustment have different scales.

The supply's output voltage is sampled by a 22kΩ/10kΩ voltage divider, with a 100nF capacitor across the upper resistor to give more feedback

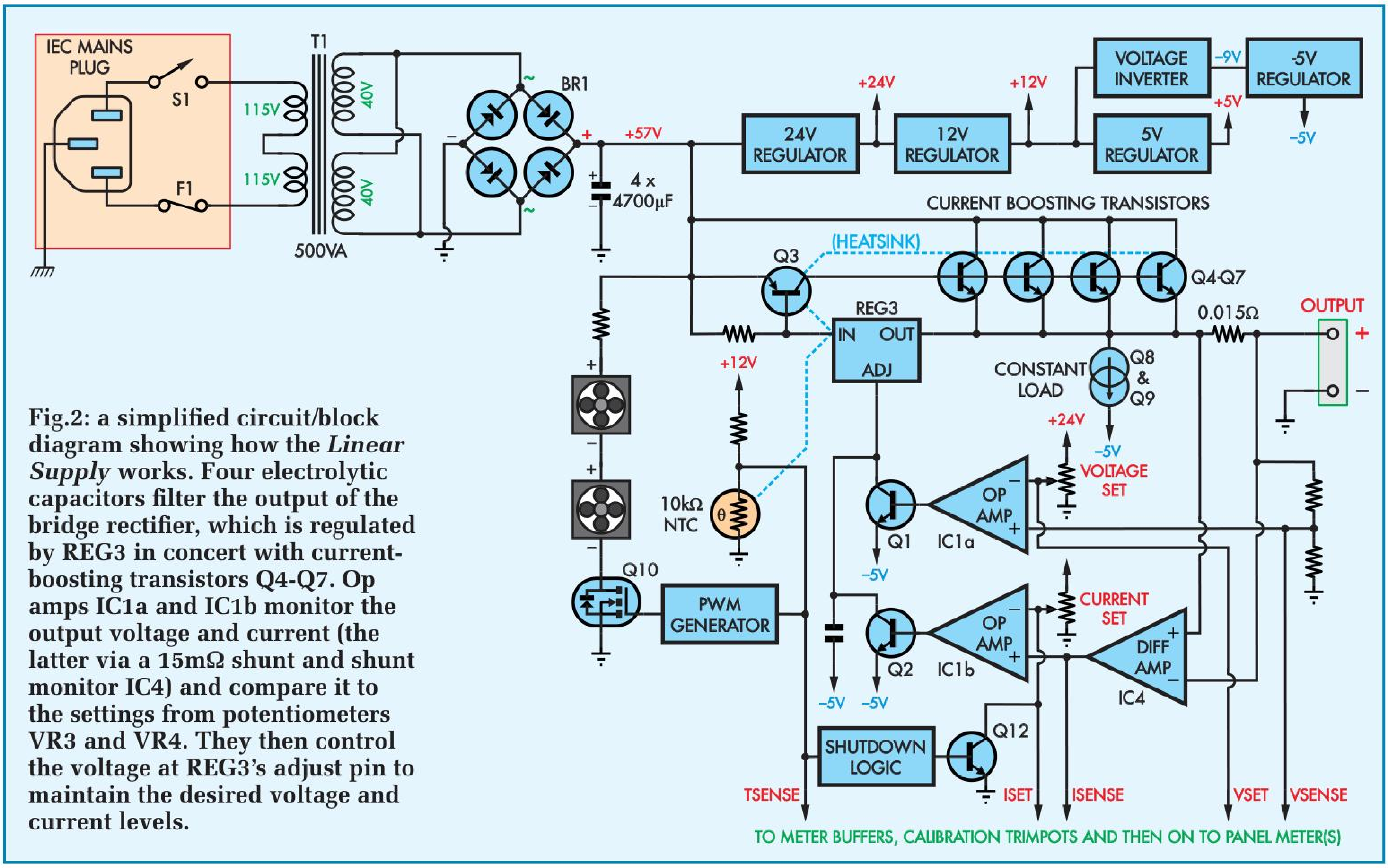


Fig.2: a simplified circuit/block diagram showing how the *Linear Supply* works. Four electrolytic capacitors filter the output of the bridge rectifier, which is regulated by REG3 in concert with current-boosting transistors Q4-Q7. Op amps IC1a and IC1b monitor the output voltage and current (the latter via a 15mΩ shunt and shunt monitor IC4) and compare it to the settings from potentiometers VR3 and VR4. They then control the voltage at REG3's adjust pin to maintain the desired voltage and current levels.

on transients, stabilising the feedback loop. The result is a 0-15.625V feedback voltage for a 0-50V output voltage.

This divider is necessary to keep the feedback voltage within the input voltage range of op amp IC1a, which runs from the 24V supply.

For the normal 0-50V output range, VR1 is adjusted to give 15.625V at TP1 with VR3 rotated fully clockwise (the voltage at TP5 should be similar). If you want to limit the voltage output to 45V, avoiding the loss of regulation at higher current settings, it can be adjusted to 14.04V instead.

Current feedback from the 15mΩ shunt is via the INA282 shunt monitor (IC4) which has a gain of 50 times.

That means that a 1A output current results in 750mV (1A × 15mΩ × 50) at output pin 5 of IC4. So at the maximum output current of 8A, we get 6V from IC4. Therefore, VR2 is adjusted to give 6V at TP3 with VR4 rotated fully clockwise (the voltage at TP6 will be similar).

Under normal operation, it is expected that TP2 (V_{SENSE}) will track TP1 (V_{SET}) as the output voltage follows the control. If current limiting is occurring, then TP4 (I_{SENSE}) will track TP3 (I_{SET}), and the voltage at TP2 will be less than TP1.

There are 100nF capacitors from the VR3 and VR4 wipers to –5V. They keep the impedance of these control lines low to minimise noise pickup, which would otherwise make its way to the supply's output.

Getting back to the control circuitry, the output from each op amp stage in IC1 (pins 1 and 7) controls NPN transistors Q1 and Q2 via two 1MΩ base-current-limiting resistors. We're using BC546s because they have a 65V rating and they can see up to about 50V on their collectors.

The LM317HV only sources about 10µA out of its ADJ pin, meaning its output can only rise by 1V per millisecond as this current must charge up the 100nF capacitor between the ADJ pin and –5V. However, Q1 and Q2 can discharge this capacitor more quickly, which is important in case the output is overloaded or short-circuited, as it means the supply's voltage can be cut quickly.

Op amp IC1 and transistors Q1 and Q2 combine to provide a phenomenal amount of gain in the control loop, which is handy to have for fast response, but needs to be carefully controlled to avoid oscillation due to overshooting. The minuscule base current through the 1MΩ resistors is one way the response of the loop has been tempered.

Another is the use of the 1nF and 100nF capacitors between the op amp inputs and outputs, which dampen what would otherwise be a sharp response to a more gradual change, thus preventing oscillation.

Power output stage

As we noted earlier, the LM317HV does not carry most of the load current. It is supplemented by four power FJA4313 power transistors, Q4-Q7. These are controlled by a 68Ω pass resistor on the LM317HV's input. As its output current rises above 10mA and the voltage across the 68Ω resistor exceeds 0.6V, Q3 switches on and so do Q4-Q7, supplementing the output current.

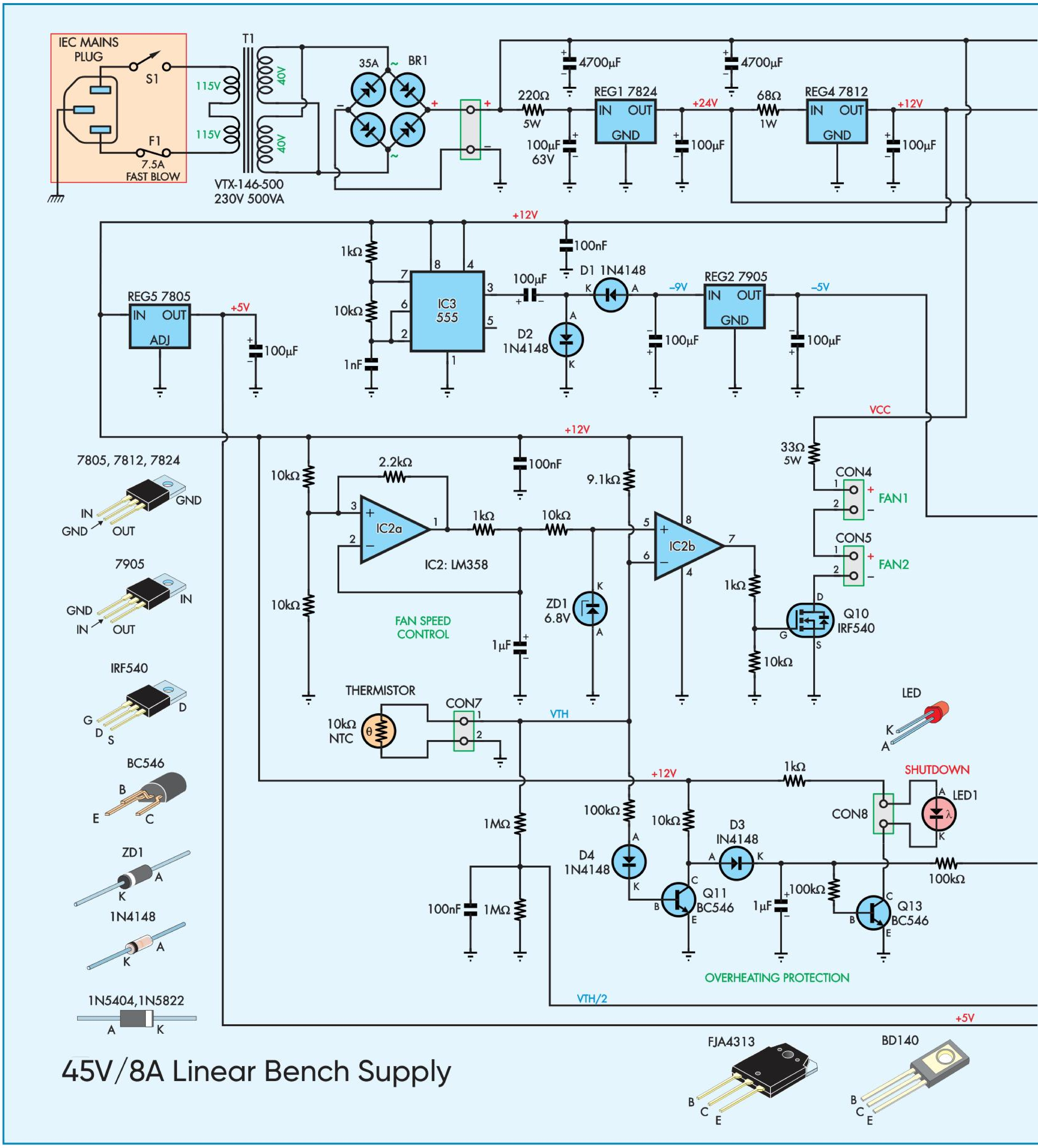
This situation is stable in that if the output current through REG1 drops due to the output transistors sourcing more current than necessary, the base current through Q3 is automatically reduced and so transistors Q4-Q7 start to switch off. Each of these transistors has a 0.1Ω emitter resistor to improve current sharing, even if the device characteristics are not identical.

At the maximum 8A output current, each of these transistors only passes about 2A, so the loss across these emitter resistors is only about 200mV.

This transistor current-booster stage again provides a tremendous amount of gain which needs to be dealt with carefully. A 100nF capacitor connects from the junction of the current sharing resistors back to the base of Q3. This provides negative feedback at high frequencies, preventing oscillation.

Transistors Q3-Q7 and REG1 (the LM317HV) are mounted on the main heatsink. As we noted, REG1 does not dissipate much power, but it is capable of thermal shutdown. It should not get hot enough for this to occur, but it does form a 'last-ditch' safeguard.

The 15mΩ high-side current shunt is monitored by IC4, an INA282 high-side shunt monitor. IC4 and the shunt


IEC MAINS PLUG
S1
F1
7.5A FAST BLOW
T1
115V
115V
40V
40V
VTX-146-500
230V 500VA
35A
BR1
4700µF
4700µF
220Ω 5W
100µF 63V
REG1 7824
IN OUT GND
+24V
100µF
68Ω 1W
REG4 7812
+12V
100µF
+12V
1kΩ
10kΩ
1nF
IC3 555
100nF
100µF
D1 1N4148
D2 1N4148
–9V
REG2 7905
–5V
100µF
100µF
REG5 7805
IN OUT ADJ
+5V
100µF
VCC
33Ω 5W
CON4
FAN1
CON5
FAN2
+12V
10kΩ
2.2kΩ
100nF
9.1kΩ
IC2a
1kΩ
10kΩ
IC2b
IC2: LM358
10kΩ
FAN SPEED CONTROL
1µF
ZD1 6.8V
1kΩ
Q10 IRF540
10kΩ
THERMISTOR
10kΩ NTC
CON7
VTH
LED
+12V
1kΩ
SHUTDOWN
LED1
CON8
100kΩ
10kΩ
D3 1N4148
1MΩ
D4 1N4148
Q11 BC546
1µF
100kΩ
Q13 BC546
100kΩ
100nF
1MΩ
OVERHEATING PROTECTION
VTH/2
+5V
7805, 7812, 7824
IN GND OUT
7905
GND IN OUT
IRF540
G D S
BC546
B E C
ZD1
A K
1N4148
A K
1N5404,1N5822
A K
FJA4313
B C E
BD140
B C E

45V/8A Linear Bench Supply

are the only two surface-mount devices used in the circuit. IC4 takes the difference between its two input voltages (the voltage across the shunt) and multiplies it by 50 before shifting it to be relative to the average voltage on its REF pins, which in this case are both connected to GND.

Thus, we have a voltage proportional to the current and referred to GND, which we can compare to the voltage on the current set potentiometer (VR4).

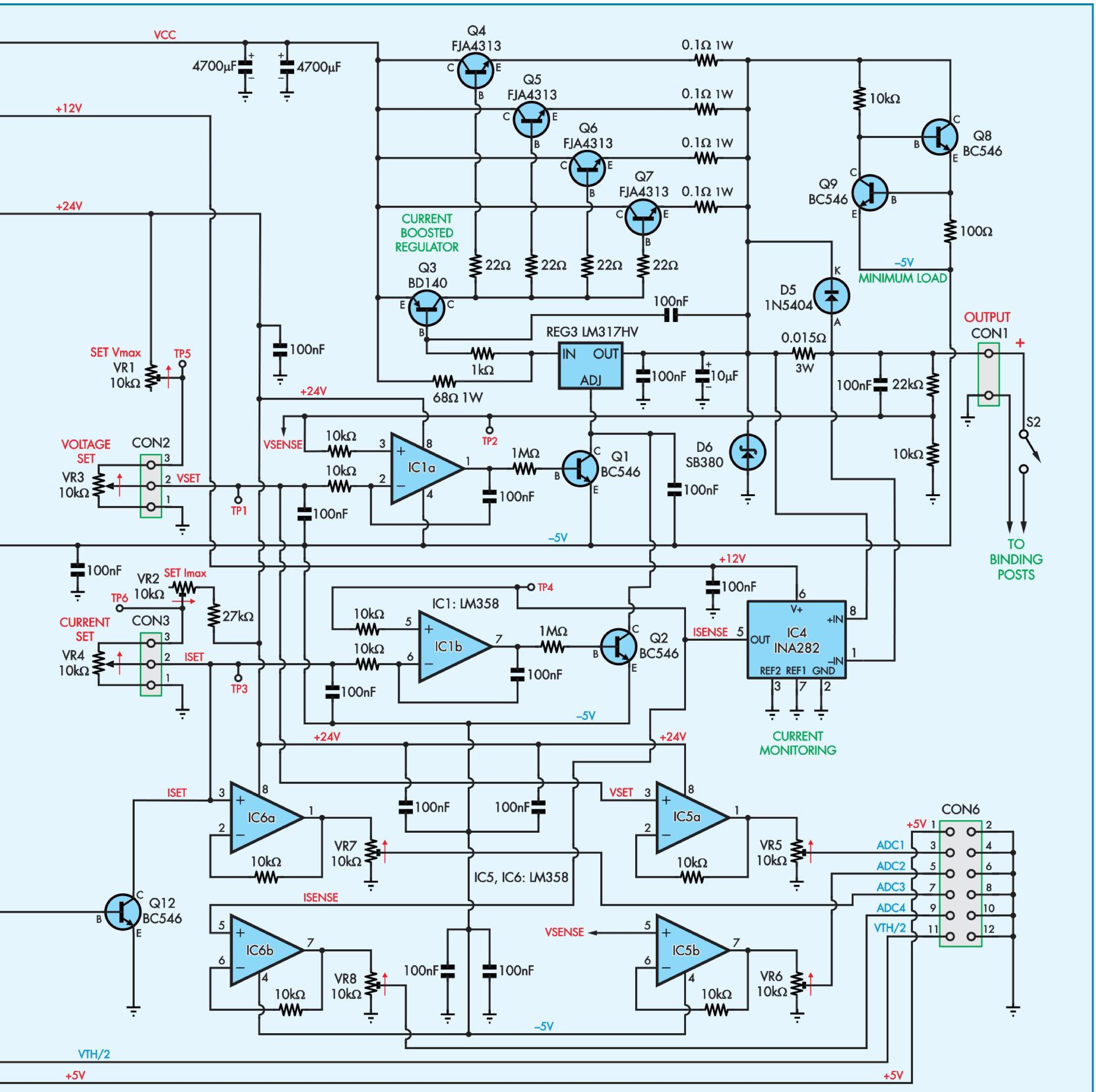
A 10µF capacitor from the output of REG3 to ground provides some smoothing and stability.

It is purposefully a small value to limit the current in case the output is short-circuited and to ensure a fast response to voltage and current changes when the supply's load is light. It's paralleled with a 100nF capacitor for better high-frequency performance.

Minimum load

The LM317HV requires a minimum output current of around 3.5mA to maintain regulation. Otherwise, the output voltage will rise.

As we cannot guarantee that there will be a load connected to the supply, we have to provide one. In a fixed voltage application, a resistor would be adequate, but not in this case.



VCC
4700µF
4700µF
+12V
+24V
Q4
FJA4313
Q5
FJA4313
Q6
FJA4313
Q7
FJA4313
C
B
E
0.1Ω 1W
0.1Ω 1W
0.1Ω 1W
0.1Ω 1W
10kΩ
Q8
BC546
Q9
BC546
100Ω
–5V
MINIMUM LOAD
CURRENT
BOOSTED
REGULATOR
22Ω
22Ω
22Ω
22Ω
Q3
BD140
100nF
REG3 LM317HV
IN
OUT
ADJ
1kΩ
68Ω 1W
100nF
10µF
D5
1N5404
K
A
0.015Ω
3W
100nF
22kΩ
OUTPUT
CON1
+
S2
10kΩ
TO
BINDING
POSTS
SET Vmax
VR1
10kΩ
TP5
100nF
+24V
VSENSE
10kΩ
10kΩ
IC1a
TP2
1MΩ
Q1
BC546
100nF
D6
SB380
100nF
VOLTAGE
SET
CON2
VR3
10kΩ
VSET
TP1
100nF
–5V
100nF
VR2
10kΩ
SET Imax
TP6
27kΩ
CURRENT
SET
CON3
VR4
10kΩ
ISET
TP3
+12V
100nF
TP4
IC1: LM358
10kΩ
10kΩ
IC1b
1MΩ
100nF
Q2
BC546
100nF
–5V
ISENSE
IC4
INA282
V+
+IN
–IN
OUT
REF2
REF1
GND
CURRENT
MONITORING
+24V
+24V
ISET
IC6a
100nF
100nF
VSET
IC5a
VR7
10kΩ
10kΩ
VR5
10kΩ
10kΩ
CON6
+5V
ADC1
ADC2
ADC3
ADC4
VTH/2
IC5, IC6: LM358
Q12
BC546
ISENSE
IC6b
VR8
10kΩ
10kΩ
100nF
100nF
VSENSE
IC5b
VR6
10kΩ
10kΩ
–5V
VTH/2
+5V
+5V

Fig.3: the full circuit of the *Linear Supply*. The regulator, control circuitry and output current monitoring are in the upper right quadrant, while the panel meter display buffer circuitry is at lower right. At centre left is the PWM fan control, with the thermal shutdown and temperature monitoring circuitry below. The mains power supply, linear regulators and negative rail generator (IC3 and D1-D2) are at upper left.

To ensure a minimum current is sunk across the full voltage range, a constant-current configuration with a pair of BC546 transistors (Q8 and Q9) is used, with the current set by a 100Ω resistor to around 6mA.

Again BC546s have been chosen to withstand the output voltage of up to 50V.

This circuit does not work unless there is more than 1.2V between its top and bottom due to the forward voltage of the two base-emitter junctions. The current is therefore sunk into the –5V rail, to ensure that regulation is maintained, even at low output voltages. At high voltages on the output, this part of the circuit can dissipate a few hundred milliwatts.

Fan control

A thermistor-controlled fan circuit is provided so that the powerful cooling fans only operate as needed. The thermistor is also used to reduce the output current in case the heatsink gets too hot, despite the fans running at full blast.

Dual op amp IC2 is powered from the 12V rail. Half (IC2a) is a triangle waveform generator, with the 1µF capacitor alternately charged and discharged between around 3V and 9V.

The triangle waveform does not have linear ramps (they're exponential), but that doesn't matter for our application. With timing components of 1kΩ and 1µF, the circuit oscillates at around 280Hz.

The triangle wave from pin 1 of IC2a is fed to the cathode of zener diode ZD1 via a 10kΩ resistor. This creates a truncated triangle wave (see Scope1), which is fed to the non-inverting input (pin 5) of the second half of the op amp, IC2b. Due to the limited current applied to ZD1, the peak voltage is around 6.5V.

The 10kΩ NTC thermistor is connected in series with a 9.1kΩ resistor, to form a voltage divider across the 12V rail. The thermistor is connected at the bottom of the divider, so that as its temperature rises, the voltage at the divider junction decreases.

At 20°C, the voltage is around 7V, dropping to around 2V at 60°C. This voltage is fed into IC2's pin 6, the inverting input. When the truncated triangle waveform voltage is above the thermistor voltage, the output pin goes high and when the triangle voltage is below the thermistor voltage, that output is low.

Thus, pin 7 of IC2b produces a square wave at 280Hz with a duty cycle that increases as the thermistor temperature increases. This drives the gate of N-channel MOSFET Q10 (IRF540) via a 1kΩ resistor, which powers the two fans. A 10kΩ pull-down on the MOSFET gate ensures it switches off when power is removed.

We have two 24V DC fans wired in series and connected via CON4 and CON5. When Q10 is on, about 9V appears across the 33Ω 5W ballast resistor, reducing the ~57V DC supply voltage to around 48-49V so they each run off about 24V. The powerful fans we have chosen draw about 280mA at 24V. If you use different fans, you will need to alter the resistor value to suit.

When the temperature at the thermistor is near ambient, the thermistor divider is at around 7V and is above the 6.5V peak set by the zener diode. Thus, the output pin of IC2b remains low, and Q10 and the fans are off.

When the divider voltage drops below the voltage set by ZD1, the fan quickly jumps up to a duty cycle of approximately 40%. This ensures that the fans start reliably, and is the reason for the presence of ZD1.

The duty cycle increases as the temperature rises until the thermistor divider voltage is below the trough of the triangle waveform, in which case Q10 and the fans are switched on 100% of the time. In this way the powerful cooling fans can dynamically respond to changes in the temperature.

Scope1-Scope4 show how the duty cycle varies in response to changes in temperature.

Thermal shutdown

The thermistor voltage is also fed to NPN transistor Q11 via a 100kΩ base resistor and also diode D4. This means that Q11 switches off if the thermistor voltage drops below 1.2V.

The high resistor value means that this part of the circuit does not affect the thermistor voltage significantly.

If the thermistor temperature rises above 80°C, then the divider voltage drops below 1.2V and Q11 switches off. Its collector voltage rises enough to allow current to flow through D3, charging the following 1µF capacitor.

This eventually provides enough base current for NPN transistors Q12 and Q13 to switch on, lighting LED1 and pulling down the wiper voltage of current set potentiometer VR4.

In practice, the current limit setpoint does not reach exactly zero when this happens, but stabilises at around 100mA, reducing the maximum dissipation in the output devices to below 10W.

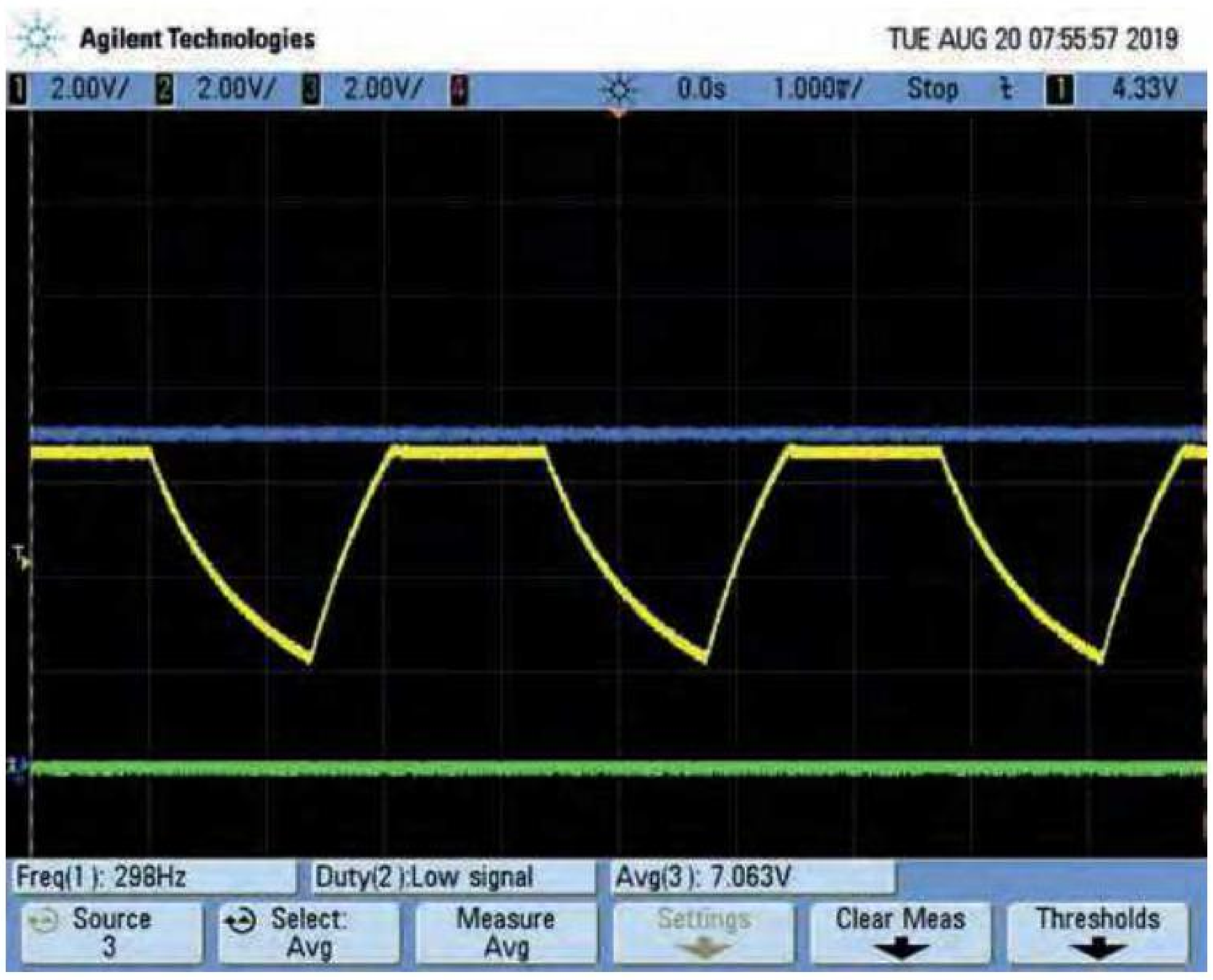
The 1µF capacitor can only discharge via the two 100kΩ base resistors, giving around a one-second delay between the thermistor voltage dropping and the current limit returning to normal.

This, in combination with the thermal mass of the heatsink, prevents the thermal limiting from switching on and off rapidly.

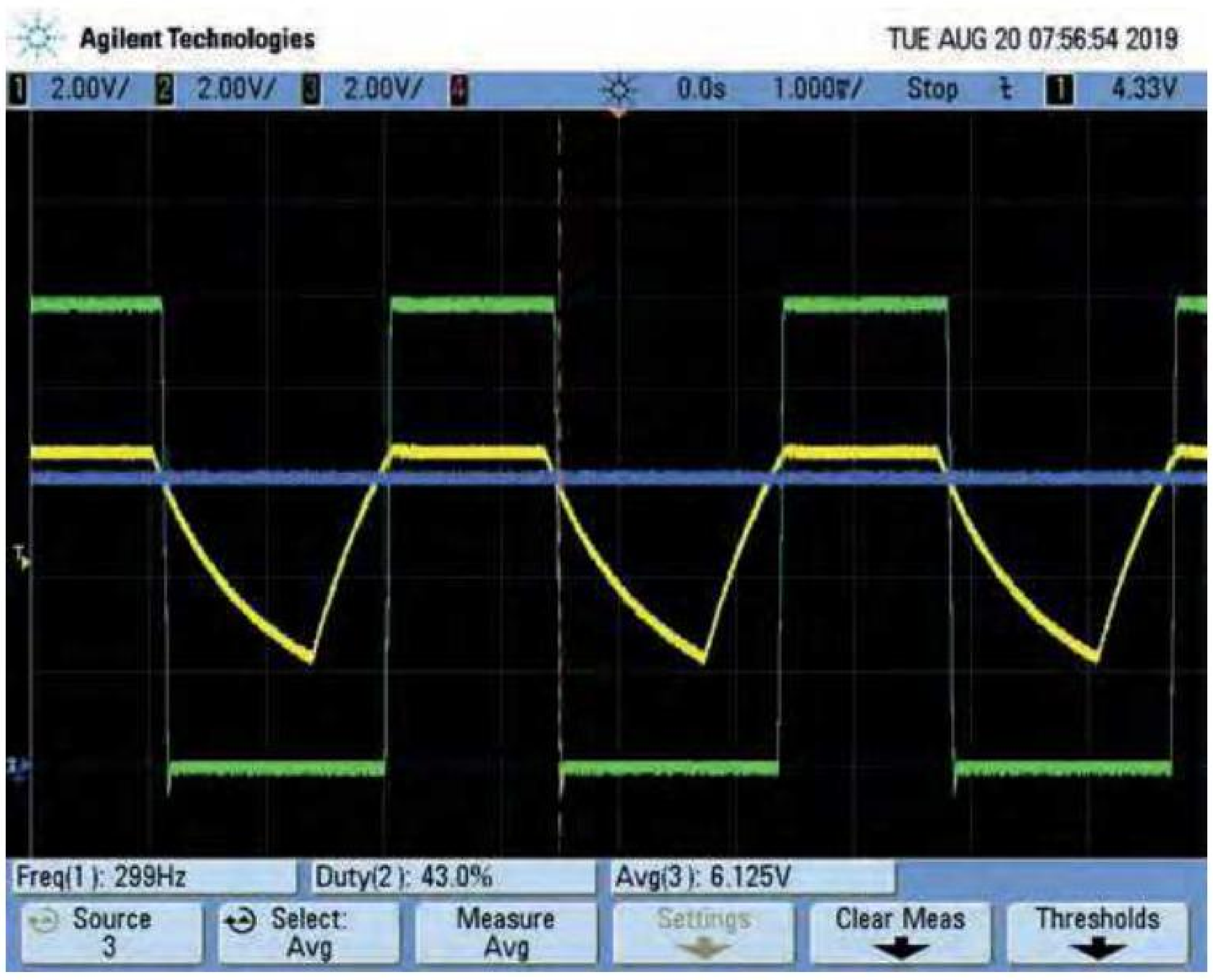
Monitoring voltages and currents

To avoid the need for hooking multiple multimeters up to the *Linear Supply* to see what it's doing, it incorporates five read-outs. These can be shown on a single LCD screen or multiple panel meters.

Regardless, the *Linear Supply* board has to provide analogue voltages to feed to these displays. These voltages are buffered by dual op amps IC5 and IC6, which are powered from the same +24V and –5V rails as control op amp IC1. They form four unity-gain amplifiers. Their non-inverting inputs are connected to TP1, TP2, TP3 and TP4.The output from each buffer is fed into a 10kΩ trimpot (VR5-VR8) to allow you to adjust the voltage scaling to suit the display(s).



Scope1: the yellow trace is the clipped 'triangle' waveform at pin 5 of IC2b, while the blue trace is the thermistor divider voltage at pin 6. Since the latter is above the former the whole time, the gate of MOSFET Q10 (green) is sitting at 0V, and so the fans are both switched off.



Scope2: the thermistor temperature has now risen enough that the divider voltage (blue) is now just below the peaks of the clipped triangle waveform (yellow) and so the gate of Q10 (green) is now a 300Hz square wave with a duty cycle of 43%. The fans are now both running at a moderate speed.

The thermal equation

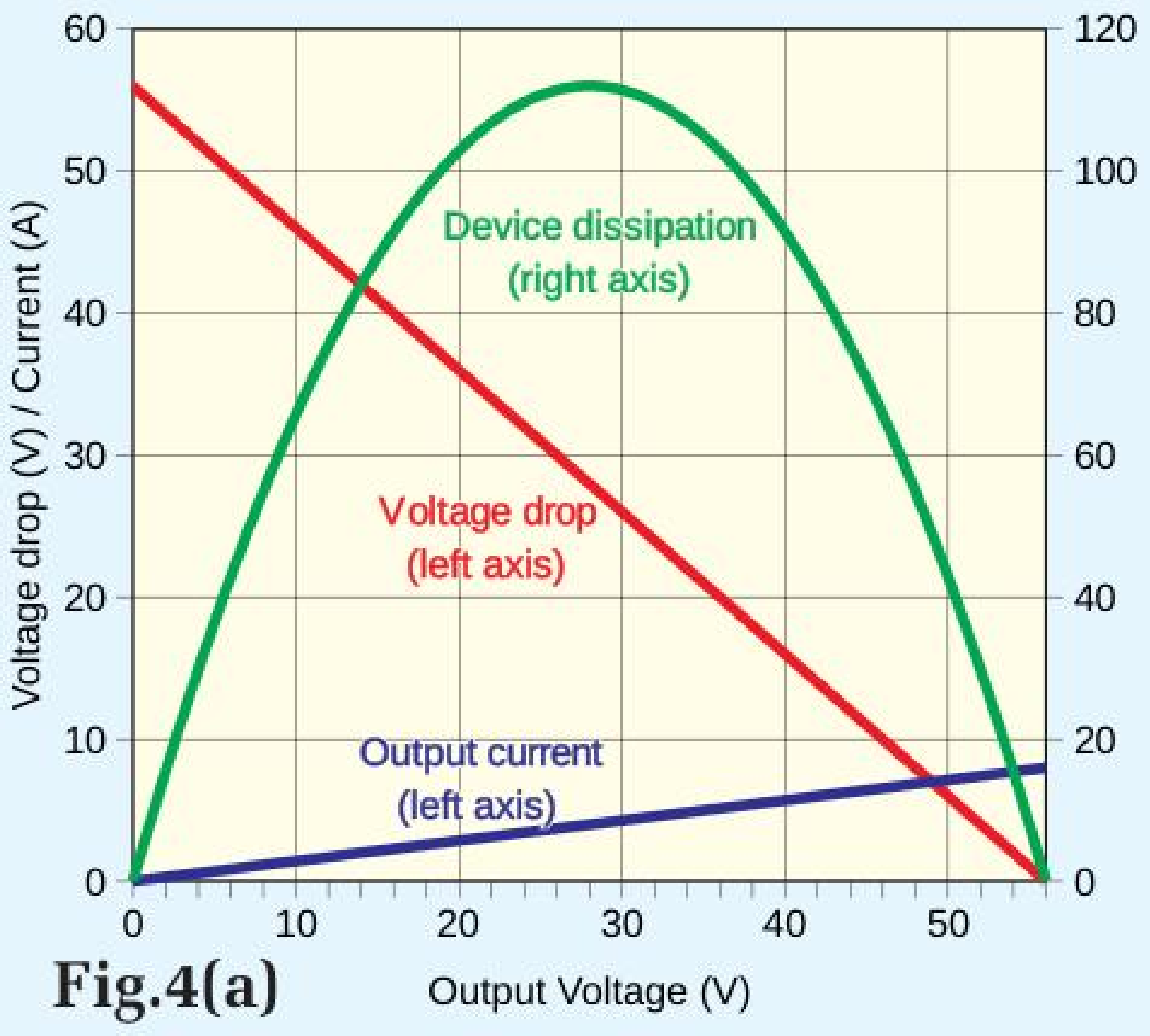


Fig.4(a)

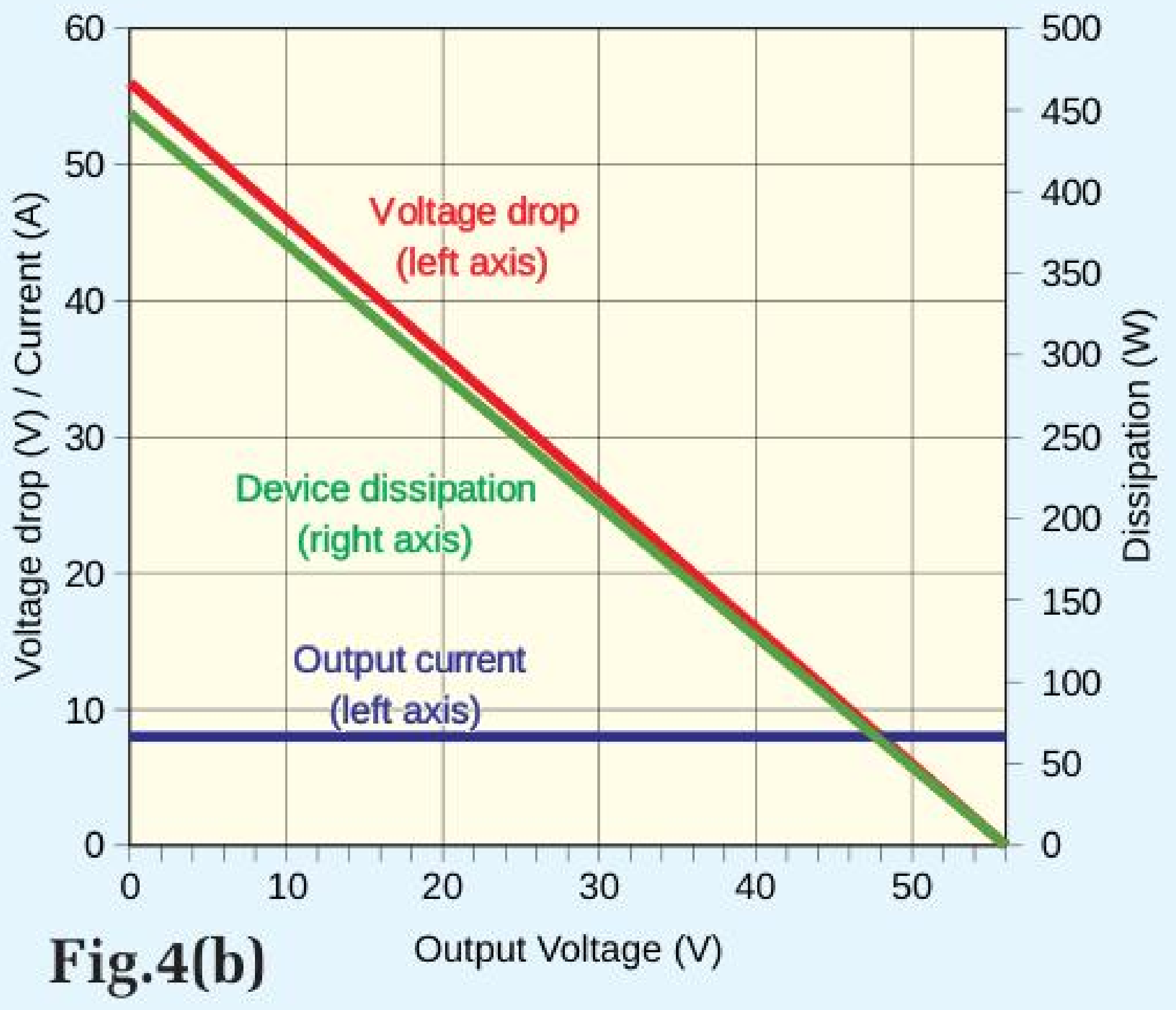


Fig.4(b)

You might notice some design parallels between this High-power Linear Supply board and a power amplifier.

Many of our power amplifiers, such as the SC200 (January-March 2018) also use a 40V transformer to provide nominal 57V rails and use four power transistors in their output stages.

While this circuit does have similarities with a power amplifier, the thermal and power considerations are significantly different.

An audio amplifier only has to deal with a relatively small load impedance variation, delivering its power into 2-10Ω or so, depending on the speaker characteristics and frequency.

The output current therefore varies more or less proportionally with the voltage. So the maximum power dissipation in the amplifier therefore occurs when the output voltage is half the supply voltage – see Fig.4(a).

On the other hand, our High-power Linear Supply cannot expect a fixed load impedance and must be capable of delivering the full load current with zero output voltage. So for the same maximum current, the maximum power is doubled, to over 400W – see Fig.4(b).

Therefore, our design needs to be able to dissipate much more power than a typical audio power amplifier module under worst-case conditions.

We initially mounted our power transistors on the heatsink using insulating pads but found that even at modest power outputs, the transistors tended to overheat, even though the heatsink was not that hot. Even switching to a thin layer of polyimide tape did not help significantly.

It was only when we directly mounted the transistors on the heatsink that we were able to keep them at a reasonable temperature when dissipating close to 100W per device.

The thermal resistance of the heatsink (with natural convection only) is quoted as 0.72°C/W, meaning that we would expect a temperature rise of 288°C above ambient with 400W total dissipation. As the maximum operating temperature of the transistors is specified as 150°C, forced – ie, with fans – cooling is necessary.

The final solution of mounting the output transistors to the heatsink, insulating it from the chassis and having two high-power fans blowing directly over its fins is necessary for correct operation of the unit under heavy load.

These trimpots effectively allow any fraction of the reference voltage to be fed to the panel meters. The thermistor voltage is scaled down by a pair of 1MΩ resistors to provide a 0-5V signal suitable for feeding to a microcontroller.

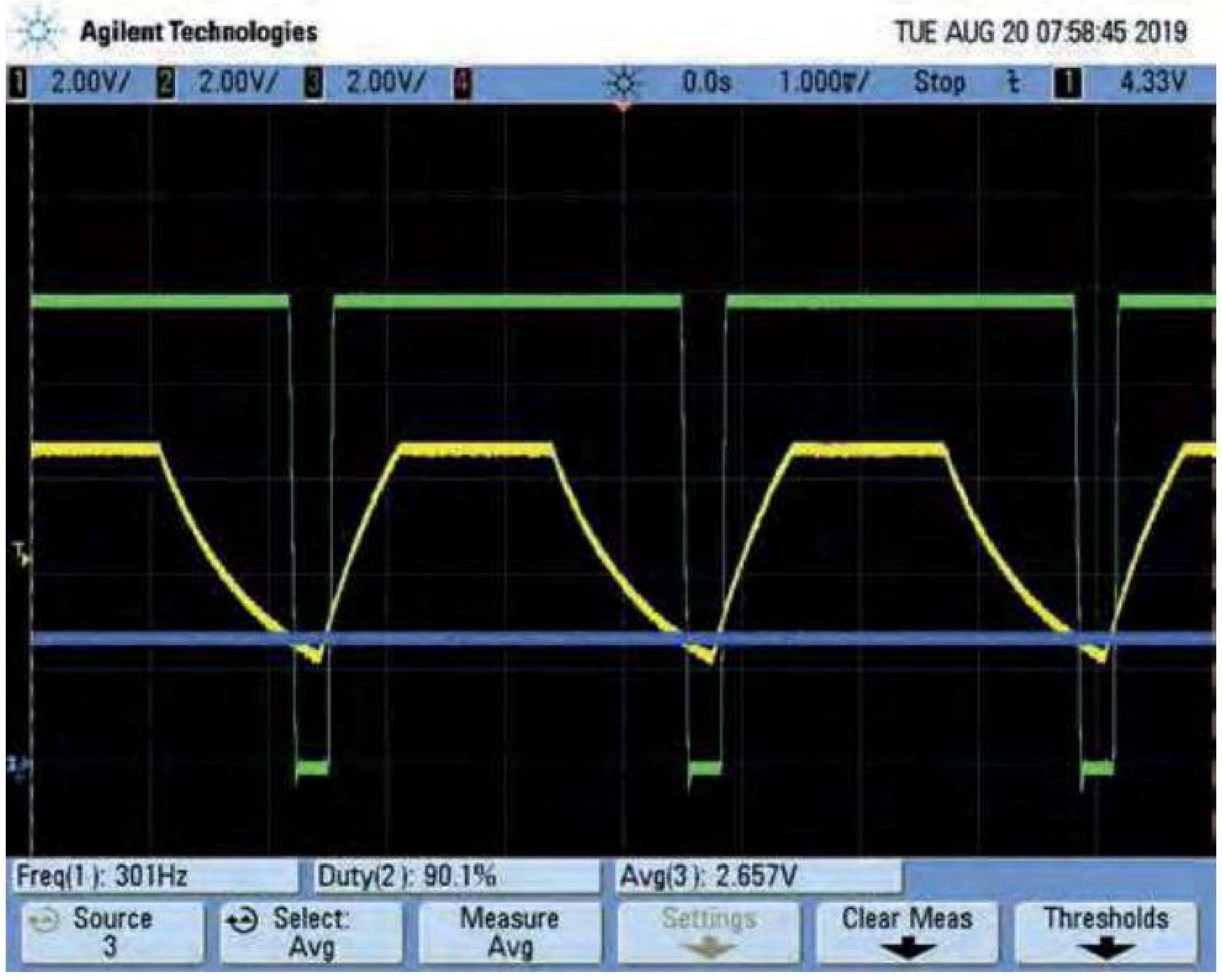
A 100nF bypass capacitor provides a low-impedance source for whatever is connected to sample it. The time constant of the 1MΩ/100nF low-pass filter is not a problem because the thermistor temperature does not change rapidly.

All the buffered signals are fed to DIL header CON6, along with ground connections and a 5V supply to run the LCD screen or panel meters.

As an example, when the *Linear Supply* is delivering 50V, there will be 15.6V at TP2. IC5b buffers this, and VR6 can be set so that 5V is fed to pin 5 of CON6 in this condition, ie, one-tenth of the actual output voltage.

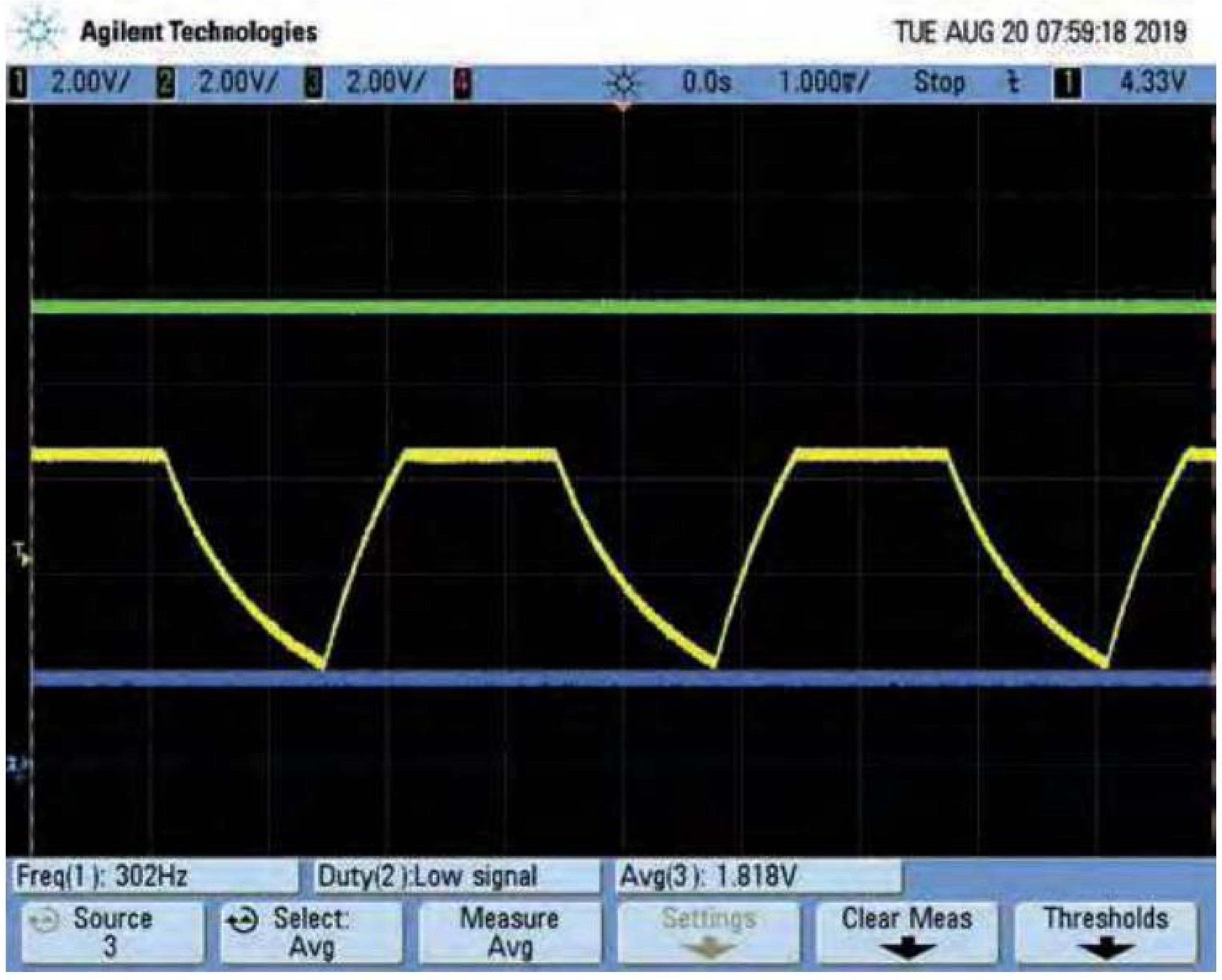
The *Linear Supply's* panel meter (see next month) just needs its decimal point set so that it reads 50.0 when receiving a 5V signal.

Similarly, the current values can be displayed on a voltmeter, with the range appropriately set by scaling and placement of the decimal point. A similar scaling by a factor of 10 is appropriate here too.



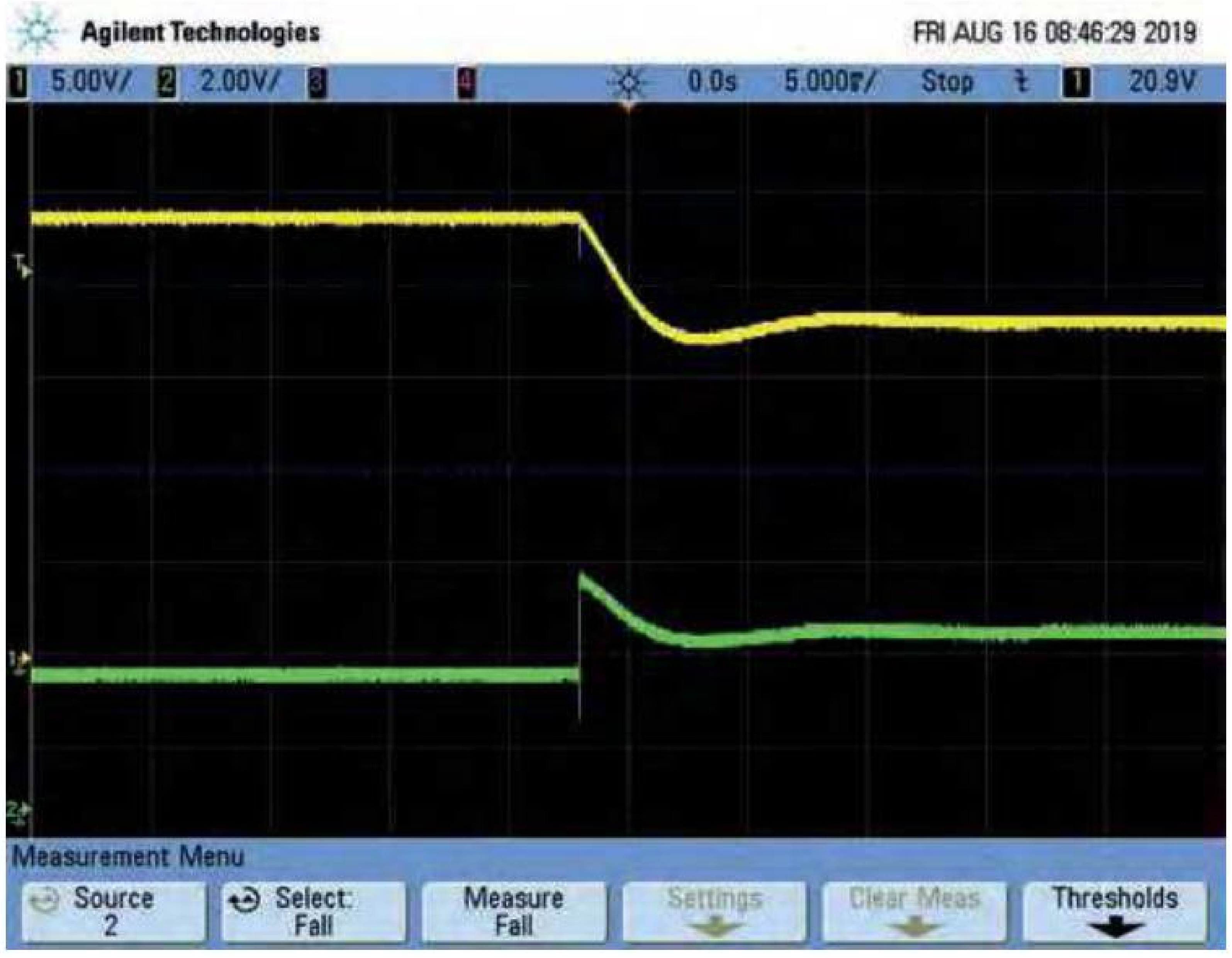
Freq(1): 301Hz
Duty(2): 90.1%
Avg(3): 2.657V

Scope3: the thermistor temperature has increased significantly, and the divider voltage (blue) has fallen, so the duty cycle at the gate of MOSFET Q10 has risen to 90%.

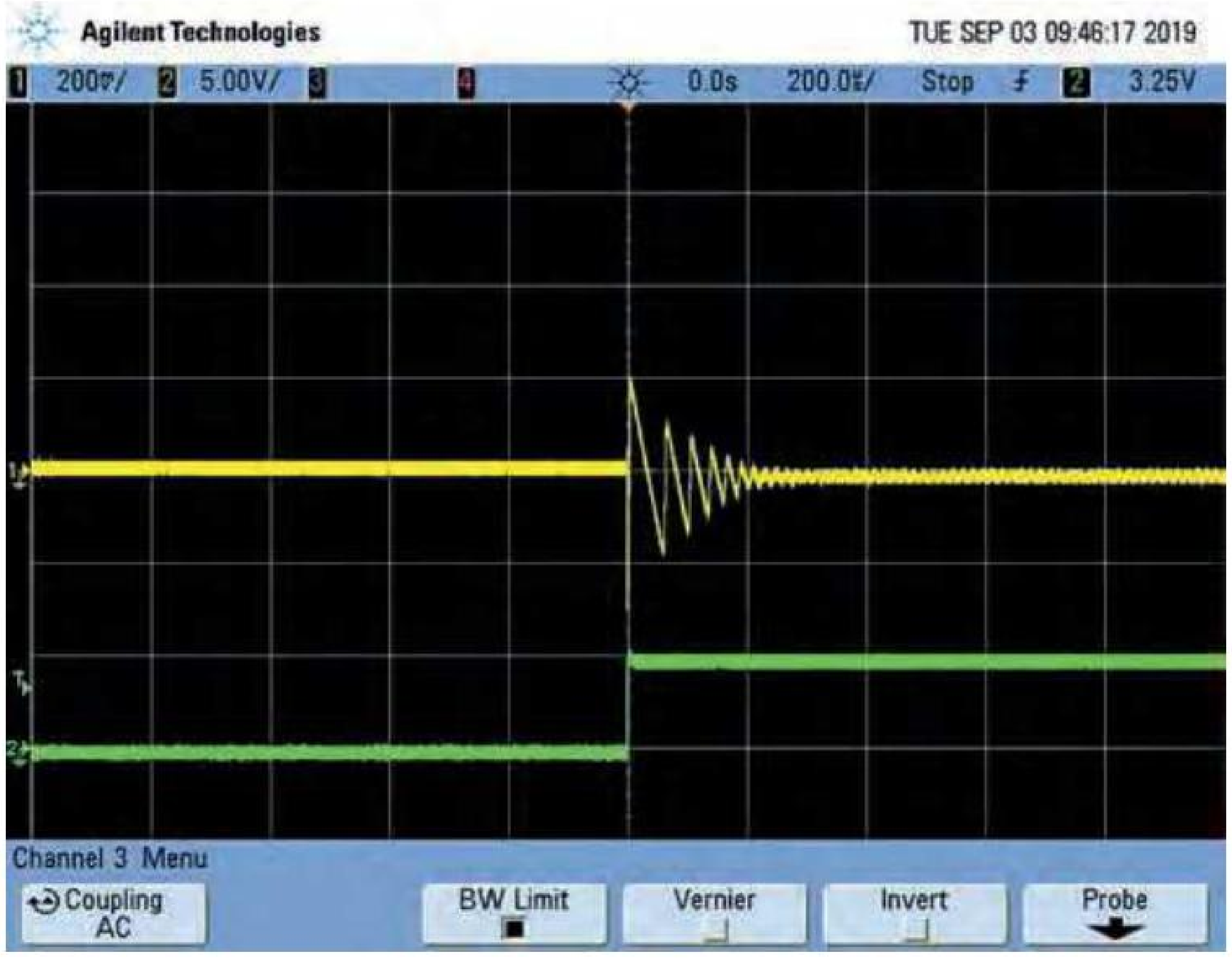


Freq(1): 302Hz
Duty(2):Low signal
Avg(3): 1.818V

Scope4: the thermistor divider voltage has now fallen further as the thermistor is very hot (above 80°C) and so the gate of MOSFET Q10 is permanently high, with the fans running continuously at full speed.



Scope5: the yellow trace shows the *Linear Supply's* output voltage, and the green trace shows its current delivery, at around 2.5A/div. It's delivering 4A at 24V into a 6Ω load, but the load impedance then suddenly drops to 3.5Ω, increasing the current to nearly 7A. The current limit has been set to around 5A, so the supply reacts within a few milliseconds to reduce the output voltage. The load current settles at the set value around 10ms later.



Scope6: this shows a 4A resistive load being connected to the *Linear Supply* while it is delivering 25V. The output is never more than 200mV from the setpoint and settles in much less than 1ms. A load with any amount of capacitance will see even less deviation than this.

Five-way Panel Meter

While we don't know of any panel meters that will be able to directly read the thermistor voltage and convert it into a temperature, our microcontroller-based *Five-way Panel Meter* design can interpret it, as well as displaying the two voltage and two current values.

The details of this low-cost *Five-way Panel Meter* will be in next month's issue, coinciding with the PCB construction and testing details for the *Supply*.

If you don't want to use that *Panel Meter* board, but you want a temperature read-out, you could feed the voltage from pin 11 of CON6 to an analogue meter and draw an appropriate scale, calibrated to match the thermistor temperature.

Internal power supply

24V linear regulator REG1 is fed from the 57V rail via a 220Ω 5W dropper (ballast) resistor. This reduces dissipation in the regulator while its 100µF input bypass capacitor prevents that resistor from affecting regulation.

The 24V rail powers the output control op amps (IC1) and the sense buffer op amps (IC5 and IC6), and is the reference voltage for the output voltage and current-adjustment potentiometers (VR3 and VR4).

The 24V rail also feeds into 12V regulator REG4 via another ballast resistor, this time 68Ω 1W. The 12V supply feeds the negative voltage generator, the current-shunt monitor IC, the thermistor and fan control, and the 5V regulator (REG5). The resulting 5V rail is for powering the panel meter/display(s).

The negative voltage generator consists of a 555 timer (IC3) operating in astable mode at around 60kHz, with a near 50% duty cycle. Its output is connected to 1N4148 diodes D1 and D2 via a 100µF capacitor, forming a charge pump.

The 100µF capacitor at pin 3 of IC3 charges up through D2 when pin 3 is high. When pin 3 goes low, D2 is reverse-biased and current instead flows through D1, charging up the 100µF capacitor at REG2's input. This results in around –9V at the input of REG2, resulting in a regulated –5V rail at its output.

Heatsinking

There are three heatsinks in this design, small flag heatsinks for the 12V and 24V regulators (REG4 and REG1) and the main heatsink for REG3, Q3-Q7 and BR1. Due to the high voltages present, regulators REG4 and REG1 have significant dissipation, despite the series ballast resistors which reduce their input voltage.

The 24V regulator is key to setting the voltage and current references, so keeping this device at a uniform temperature will help with the stability of the output.

As mentioned earlier, to efficiently get heat out of transistors Q4-Q7, they are not insulated from the main heatsink and it is therefore at around +57V DC potential. 57V DC is considered 'low voltage', but of course there are also mains voltage present around the transformer, so it doesn't hurt to use caution while working on the supply when it's powered.

The LM317HV regulator has a live tab connected to its output, which can vary anywhere between 0V and near the DC rail voltage, so it must be insulated from the main heatsink. We used a silicone pad and an insulating bush.

Similarly, the tab of Q3 is connected to its collector. If the collector were connected to the DC rail, then the output transistors would turn on hard, so this must be avoided. It too is mounted with a silicone pad and insulated bush.

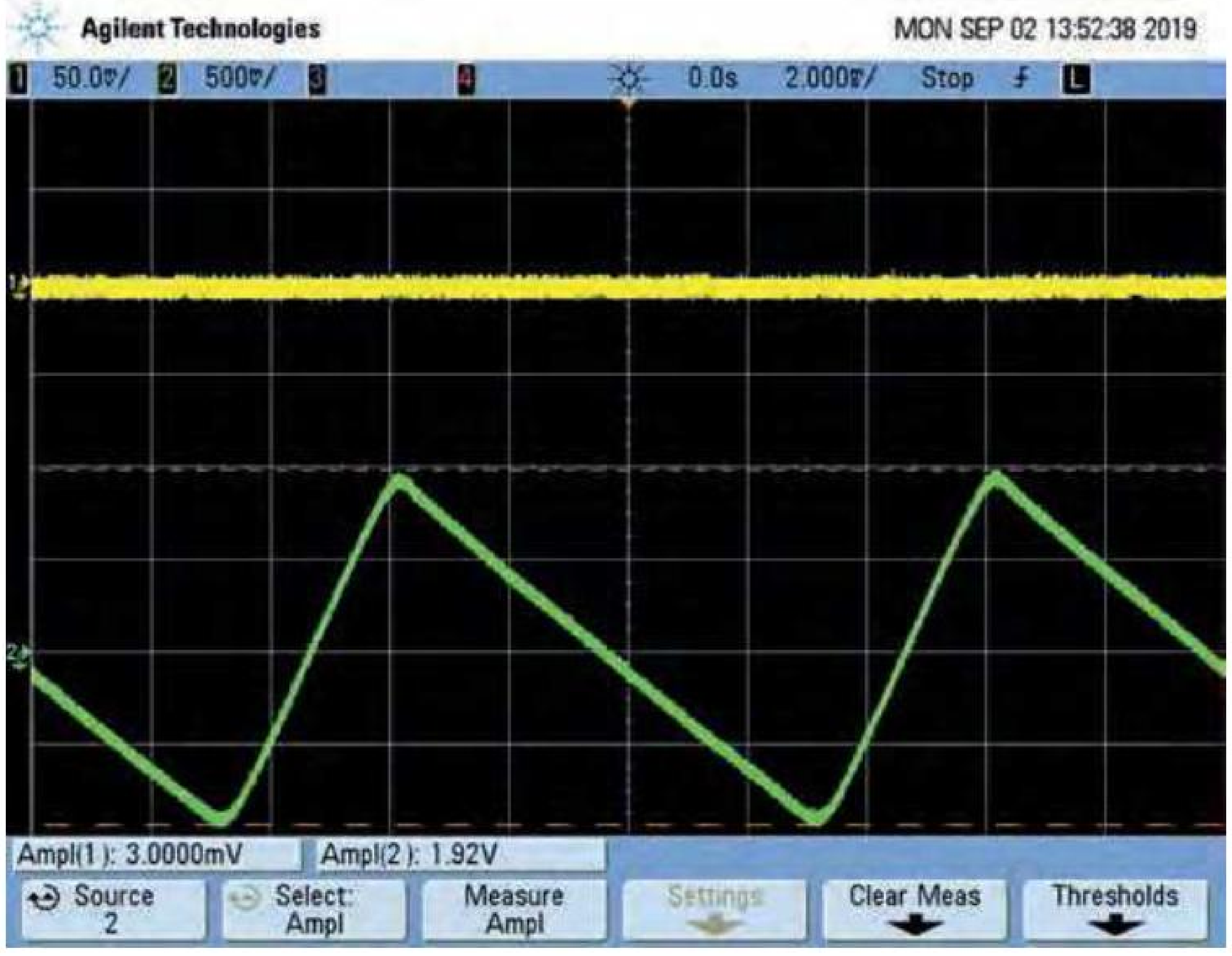
We have purposefully mounted Q3 reversed on the PCB, with its pin 3 on the left, so that its metal tab faces away from the heatsink. That's because, despite an insulating washer, we found it was still shorting to the heatsink via the screw.

Reversing the device solved that. Its dissipation is not that high, so the added thermal resistance is not a big problem. Of course, the thermistor is also mounted on the heatsink and must be insulated too. We used a stud-type thermistor, which has the active element potted, so that is already taken care of.

Performance

Scope grabs Scope5-10 demonstrate some of the performance characteristics of the *Linear Supply*. These grabs demonstrate the effects of sudden 'step' changes in the operating conditions. In reality, most changes won't occur so suddenly.

Do note that the *Linear Supply* can respond quickly to changes in load without excessive overshoots, including switching into current limiting when necessary. The scope grabs demonstrate that it typically responds within milliseconds to these sort of changes. See the details of the individual tests underneath the scope grabs.


Agilent Technologies MON SEP 02 13:52:38 2019

1 50.0mV/ 2 500mV/ 3 4 0.0s 2.000ms/ Stop

Ampl(1): 3.0000mV Ampl(2): 1.92V

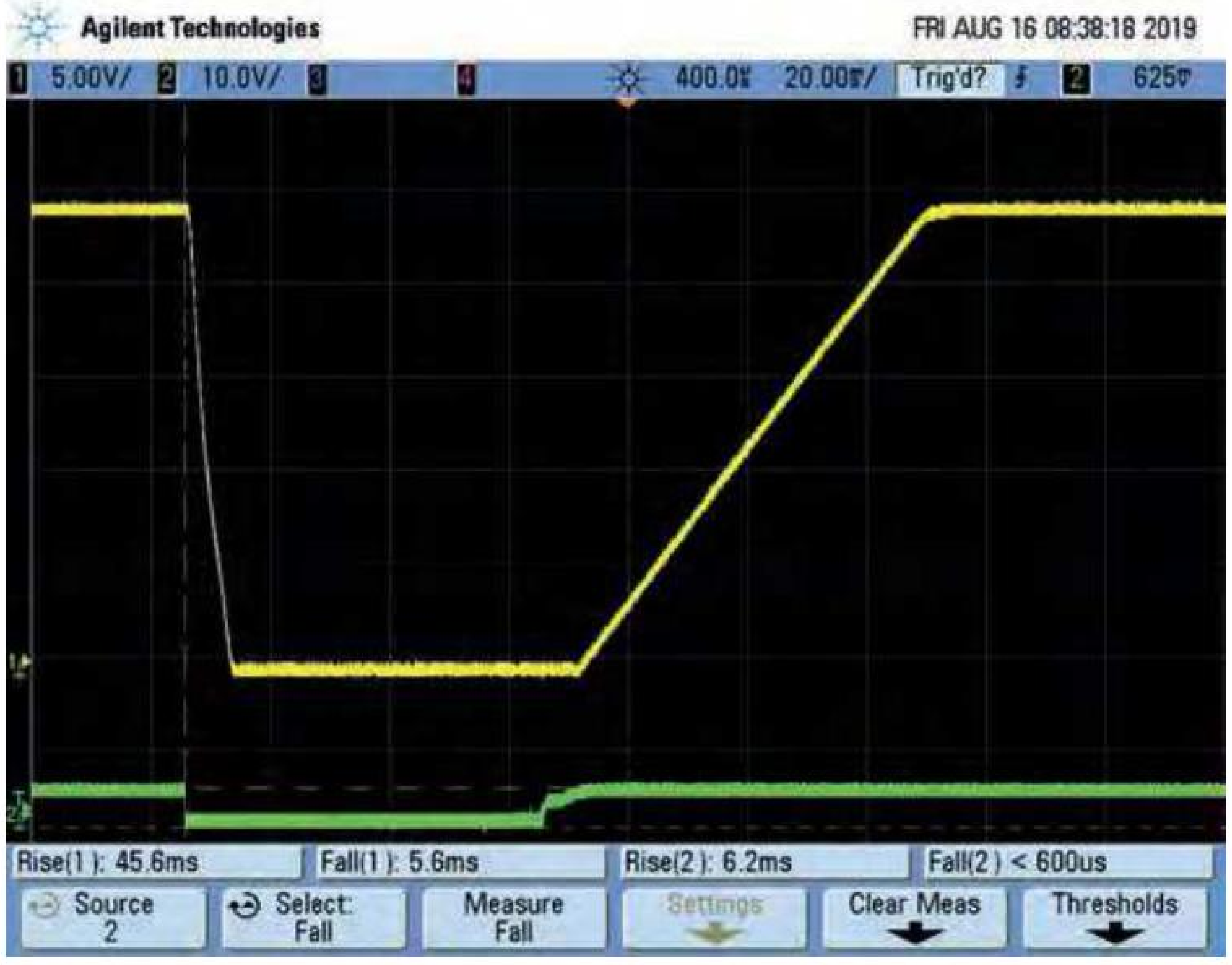
Scope7: the green trace shows around 2V of ripple on the pre-regulator 4 × 4700µF capacitor bank with the *Linear Supply* delivering 4A into 25V. The yellow trace is the *Supply's* output. The scope measures 3mV of ripple, but this is comparable in magnitude to the noise that the scope probes pick up when grounded.

We also did some thermal tests to determine how well the *Linear Supply* handles heat dissipation. As noted in our panel about 'The Thermal Equation', the *Linear Supply* works hardest when the output voltage is low, but the current is high. In these cases, the full supply voltage appears across the output transistors.

For example, dumping 8A into a short circuit means that the *Linear Supply* is delivering around 400W into the heatsink. During our scope grab tests, at 25V and 4A, it is dissipating around 100W.

Under the latter condition, the thermistor registers around 20°C above ambient, and the fans run at around half speed.

One of our more severe tests involved connecting a 2Ω dummy load. With the output set to 8A, the voltage reaches 16V, and the *Linear Supply* is dissipating around 300W. Under these conditions, the thermistor reached 77°C (around 55°C above ambient) after around 10 minutes and then held steady.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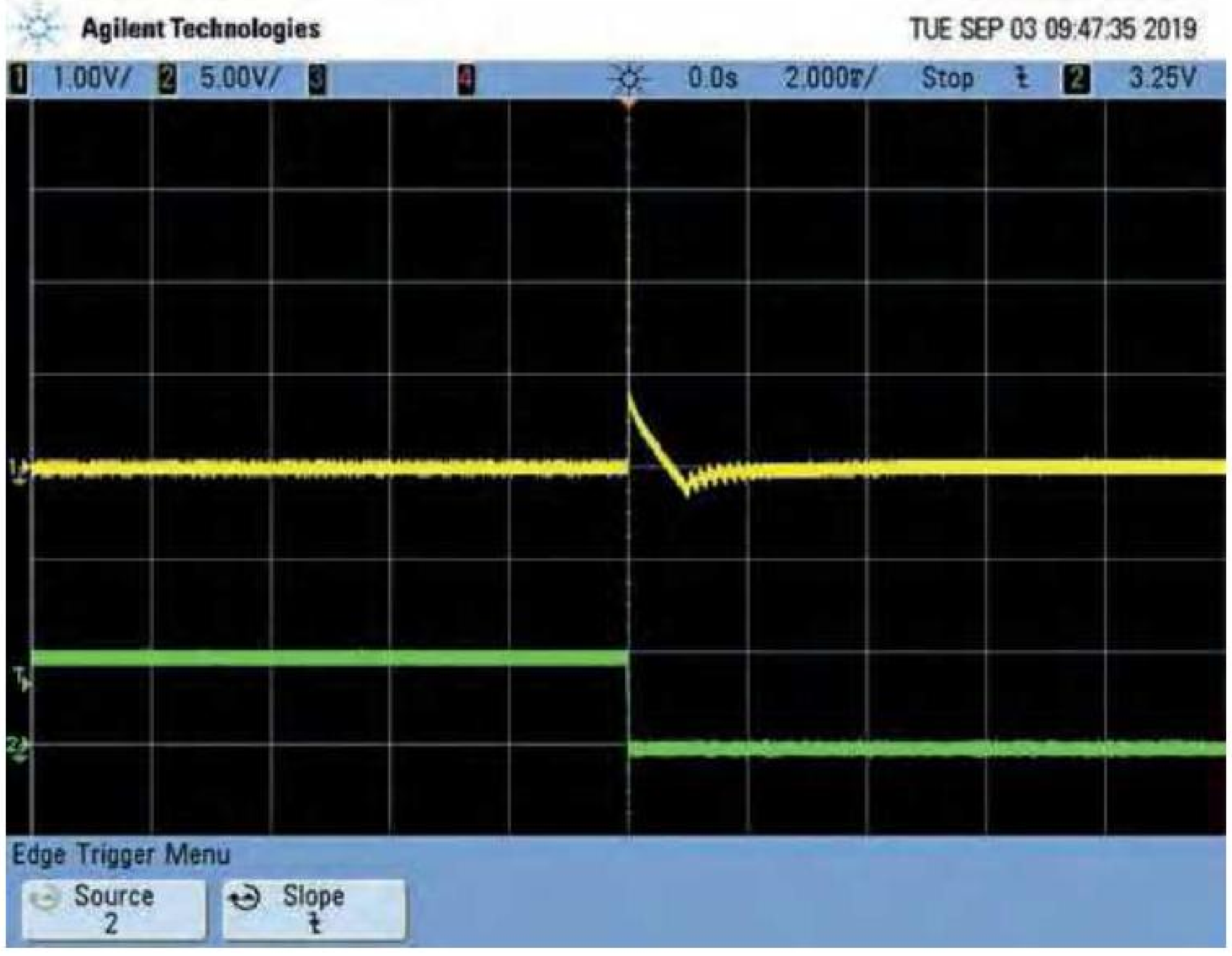
1 5.00V/ 2 10.0V/ 3 4 400.0µs 20.00ms/ Trig'd? 2 625mV

Rise(1): 45.6ms Fall(1): 5.6ms Rise(2): 6.2ms Fall(2) < 600us

Scope9: here we have simulated a step-change in the voltage control input by shorting the V_{SET} point to ground and then releasing it. The output voltage drop is much quicker than the rise, ensuring that the chance of overshoot is minimised under dynamic conditions.


Agilent Technologies TUE SEP 03 09:47:35 2019

1 1.00V/ 2 5.00V/ 3 4 0.0s 2.000ms/ Stop 2 3.25V

Scope8: This is the reverse of the scenario seen in Scope6, with a 4A resistive load being disconnected from the Supply at 25V. There is around half a volt of overshoot followed by a lesser amount of undershoot and the output settles completely within 2ms.

Contrary to what you might think, delivering 45V at 8A is not that stressful to the supply, as there is only about 10V across the output devices and thus a dissipation of around 80W.

Delivering 8A into a short circuit is more difficult; the supply can manage this, but only for a few minutes at a time before it enters thermal current limiting.

Next month

As promised earlier in this article, our November issue will commence the full construction details, including the parts list. If you want to be sure not to miss that issue, why not subscribe to *Practical Electronics?* (See page 4).

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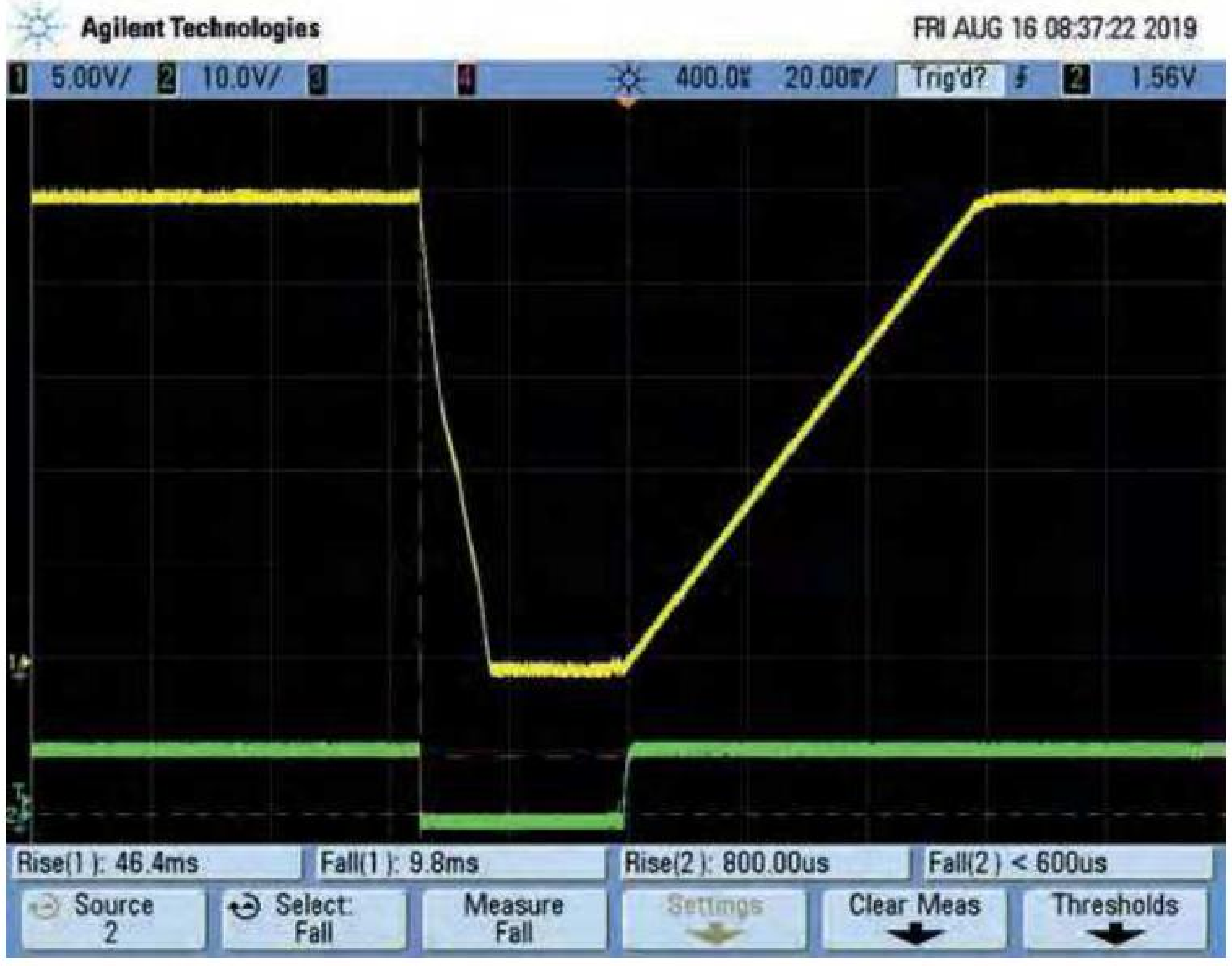
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1 5.00V/ 2 10.0V/ 3 4 400.0µs 20.00ms/ Trig'd? 2 1.56V

Rise(1): 46.4ms Fall(1): 9.8ms Rise(2): 800.00us Fall(2) < 600us

Scope10: this current control step-change test shows a similar response as in Scope9. Again, the fall is faster, indicating that the *Linear Supply* is designed to respond to over-current conditions quickly. Note that there there is no visible overshoot.

Practical Electronics

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For those times when near enough isn't good enough!

There's a 'law' in electronics, which says you can never have too much test equipment. Even if it is pretty specialised; even if you only need it once every blue moon, there will come a time when you do need it! This one fits that bill perfectly – you're not going to need it every day, but when you do, you'll thank your good sense that you do have one on hand!

By Jim Rowe

So what are we talking about? It's a *Precision 'Audio' Signal Amplifier*. It's used when you need to know – exactly – what 'audio' signal you're dealing with.

'Audio' is in quotes because it will actually handle signals way below the normal audio range (down to just 6Hz) and it can go all the way up to more than 230kHz… or even higher.

Talk about the proverbial 'DC to Daylight' amplifier… this one is not far off!

It can deliver a particularly impressive 30V peak-to-peak (10.6V RMS) up to around 230kHz. It has two switched gain settings of either 1.00 times (0dB) or 10.000 times (+20dB), and it's powered from a standard 12V AC plugpack.

OK, so why would you want one?

Let's say you've built some audio gear and want to check it out properly. Or maybe, you need to accurately calibrate other test gear. Or perhaps (and we imagine this will be the biggest market) you're in the service game and need to troubleshoot a misbehaving unit.

You may already have a low-cost waveform generator (there are many on the market these days) and they are increasingly built into DSOs.

They can usually generate sine, square, triangle and often 'arbitrary' waveforms with varying frequencies and amplitudes.

But the maximum amplitude is usually limited to about 5V peak-to-peak and often, that simply isn't sufficient.

This project overcomes that limitation. It can be connected to the output of the waveform generator to provide exactly ten times gain, boosting the signal level to just over 10V RMS.

And because the gain applied is very precise, you don't need to check the output level. You just set the generator to produce a waveform with 1/10th the needed amplitude and the signal amplifier does the rest.

I came up against this problem when calibrating our new *Digital Audio Millivoltmeter*, described starting on page 32 of this issue.

I have a few signal generators, but none of them could produce a sinewave with sufficient amplitude to calibrate its 'HIGH' range. So I decided to design and build this precision amplifier, to generate accurate signals of a high enough amplitude for me to calibrate it.

And I realised just how useful this would be for other audio projects! It provides a choice of two accurately known gain ranges (1:1/0dB or 1:10/+20dB), over a relatively wide range of frequencies, from 20Hz up to beyond 200kHz.

Circuit details

The *Signal 'Audio' Amplifier* circuit is shown in Fig.2. As you can see, the amplifier itself (lower section) is quite straightforward.

That's because we are using a rather special op amp, the Analog Devices ADA4625-1 (IC1). It offers very high input resistance, thanks to the use of

JFET input transistors. It has a typical gain-bandwidth product of 18MHz, very low noise, fast settling time (to within 0.01% in 700ns), a rail-to-rail output swing and the ability to operate from a 'single supply' of up to 36V.

Other features include a low output resistance in closed-loop mode (typically 2Ω when gain=1 or 18Ω when gain=10) and the ability to drive load capacitances up to 1nF in closed-loop unity-gain operation.

We are using IC1 in a standard non-inverting configuration, with the input signal from CON1 coupled to its non-inverting input (pin 3) via a 1μF metallised polyester capacitor.

The output from IC1 is then fed to output connectors CON2 and CON3 via another 1μF metallised polyester coupling cap, with a 51Ω protective (and impedance-matching) resistor in series with each connector.

Switch S1 is used to alter the feedback around IC1, to provide either unity gain or a gain of 10.

In the A=10 position, the 100kΩ 0.1% resistor forms the top arm of the feedback divider, while the lower arm is formed by the series combination of the 10kΩ and 820Ω fixed resistors, together with VR1, a 15-turn 500Ω trimpot.

The trimpot allows us to set the amplifier's gain to exactly 10.000, by compensating for within-tolerance variations in the value of the 10kΩ and 820Ω resistors (both 1% tolerance) as well as the 100kΩ 0.1% tolerance resistor.

Although it's easy to calculate the nominal lower-arm resistance for a gain of 10.000 (it's 11.111kΩ), this would need to be made up from at least two more 0.1% tolerance resistors (11.0kΩ and 110Ω), to give a gain of 10.0009 with a tolerance of +0.018% and –0.0162%.

By using two 1% tolerance resistors and a 15-turn trimpot, we can achieve even better potential accuracy at a significantly lower cost.

Features and specifications

Input impedance	100kΩ//9pF
Output impedance	51Ω (each output)
Gain	A=1 (0dB) or A=10 (+20dB)
Frequency range (see Fig.1)	For A=1: 6Hz to >1MHz (+0,-0.3dB)
	For A=10: 20Hz to 230kHz (+0,-0.3dB)
Maximum input signal level	For A=1: 10.6V RMS (+20.5dBV)
	For A=10: 1.06V RMS (+0.5dBV)
Maximum output signal level	10.6V RMS (+20.5dBV)
THD+N	For A=1: 0.0007% (-103dB)
	For A=10: 0.007% (-83dB)
Power supply	12V AC at <100mA

But how do you set the gain to exactly 10.000? You just need a relatively accurate DMM. You measure the value of the 100kΩ 0.1% resistor (which should be between 99.9kΩ and 100.1kΩ), then divide that by nine, and adjust VR1 so that the total lower-arm resistance matches the calculated value (which should be close to 11.111kΩ).

The upper part of the circuit exists primarily to generate a 32V DC supply voltage from the 12V AC plugpack, so that IC1 can deliver output signal amplitudes as high as 30V peak-to-peak or 10.6V RMS.

This is achieved in two stages. First, diodes D1 and D2 and the two 470μF capacitors form a simple 'voltage doubler' rectifier configuration, which derives about 38V DC from the incoming 12V AC. This is followed by voltage regulator REG1, an SMD version of the familiar LM317 adjustable regulator. Here it's configured to provide a regulated output of 32V which is fed to IC1 via a 100Ω resistor.

The 220nF capacitor from the ADJ (Adjust) pin to ground improves its ripple rejection, which is helpful here because with the voltage doubler configuration, each filter capacitor only recharges at 50Hz.

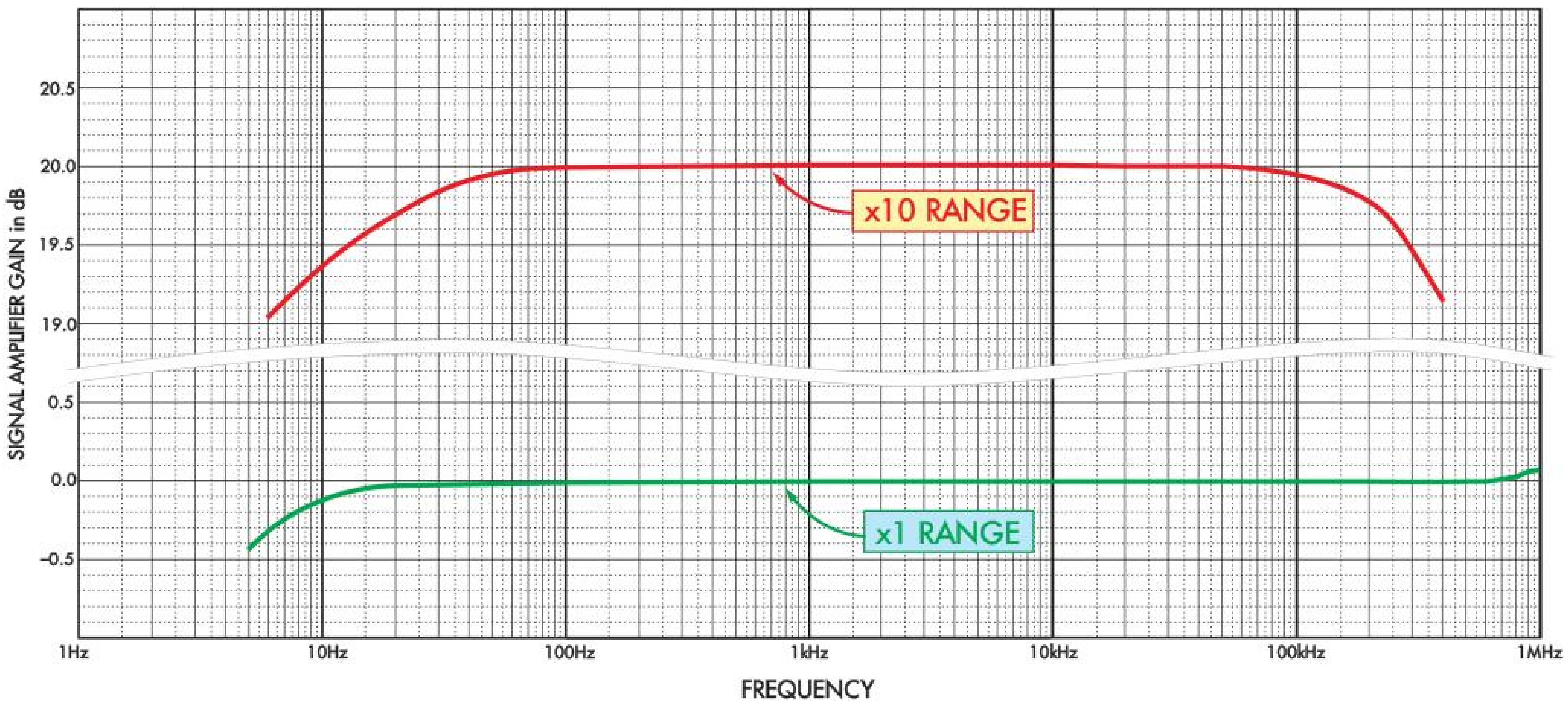


Fig.1: this shows a frequency response plot for the *Signal Amplifier* at both gain settings. The response in both modes is entirely flat from 100Hz to 50kHz, so ideally, calibration and measurements should be made within that range. But it gives acceptable performance (within 0.3dB) from 20Hz to 230kHz, which more than covers the audio range.

The 32V rail is also used to provide a 'half supply voltage' bias of 16V for the non-inverting input of IC1, via a 10kΩ/10kΩ resistive divider with a 220nF ripple filter capacitor.

LED1 is a power-on indicator, connected to the +32V line via a 3.3kΩ series resistor.

Construction

Almost all of the circuitry and components are mounted on a single PCB which fits inside a diecast aluminium box, for shielding.

The PCB measures 92 × 51mm and is coded 04107191. Refer now to the overlay diagram (Fig.3) along with the matching photo.

The only components not mounted on the PCB are input and output connectors CON1-CON3 and range selection switch S1. These all mount on the box lid/front panel, with short lengths of hookup wire linking them to the PCB.

It's easiest to fit the SMD components to the PCB first, starting with the passives (resistors and capacitors) and then REG1 and IC1.

Make sure IC1's pin 1 dot/divot (or bevelled edge) is oriented as shown in Fig.3.

Then fit the leaded parts, starting with the 100kΩ 0.1% resistor and diodes D1-D4 (with the orientations as shown), trimpot VR1, the two 1μF capacitors, the two 470μF and 47μF electrolytic capacitors (longer lead towards + sign) and then the power input connector, CON4.

The final step is to fit LED1, which is mounted vertically just below the centre of the PCB. First solder a 2-pin SIL header to the PCB, then solder the LED's leads to the header pins, with the LED anode towards the front.

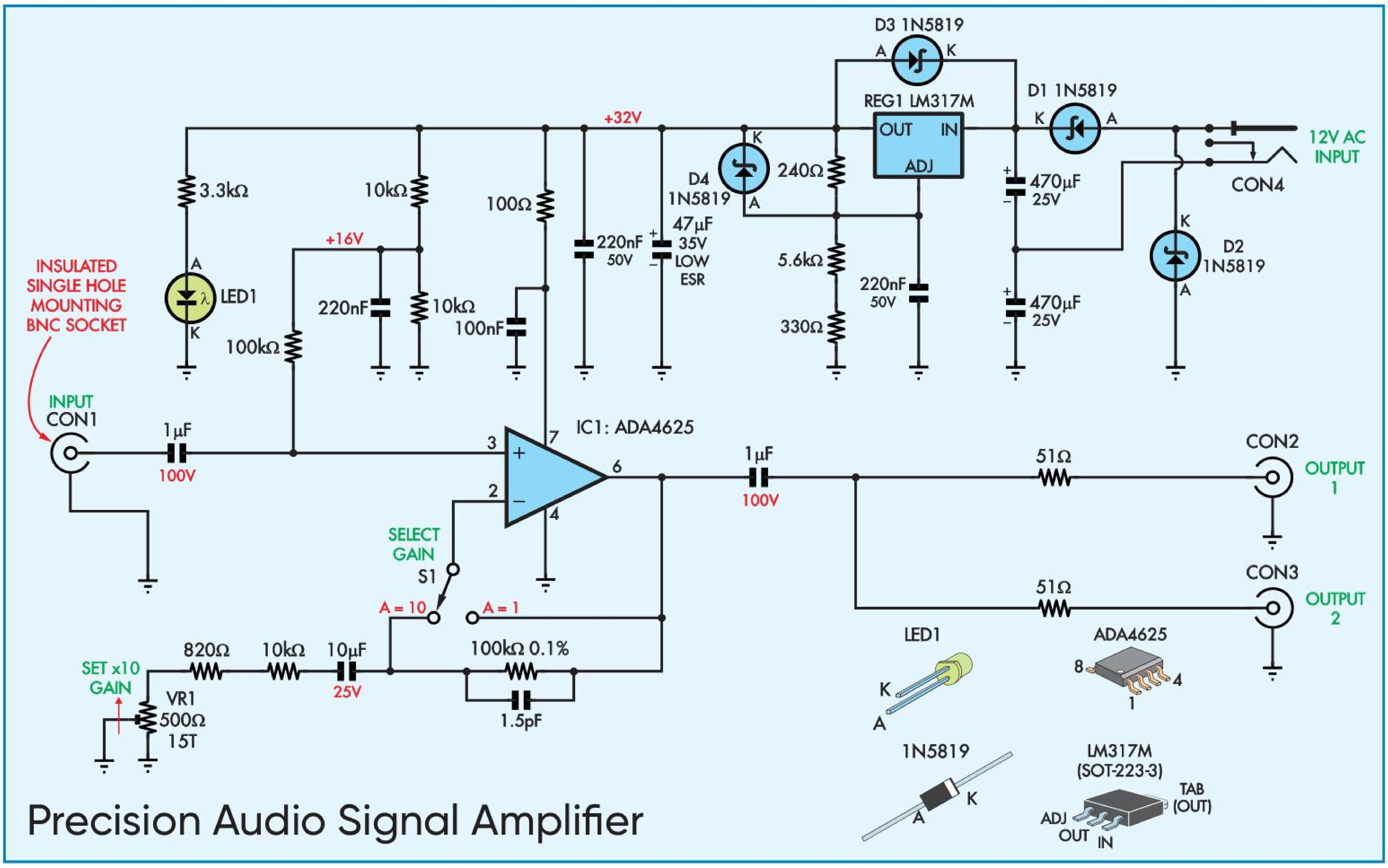

D3 1N5819
A
K
REG1 LM317M
OUT
IN
ADJ
D1 1N5819
K
A
+32V
12V AC INPUT
CON4
3.3kΩ
10kΩ
100Ω
D4 1N5819
K
A
240Ω
470μF 25V
+16V
47μF 35V LOW ESR
220nF 50V
5.6kΩ
220nF 50V
D2 1N5819
K
A
INSULATED SINGLE HOLE MOUNTING BNC SOCKET
LED1
A
K
220nF
10kΩ
100nF
330Ω
470μF 25V
100kΩ
INPUT CON1
1μF 100V
IC1: ADA4625
3
7
2
6
4
1μF 100V
51Ω
CON2
OUTPUT 1
SELECT GAIN S1
A = 10
A = 1
51Ω
CON3
OUTPUT 2
820Ω
10kΩ
10μF 25V
100kΩ 0.1%
1.5pF
SET x10 GAIN
VR1 500Ω 15T
LED1
K
A
ADA4625
8
4
1
1N5819
A
K
LM317M (SOT-223-3)
TAB (OUT)
ADJ
OUT
IN
Precision Audio Signal Amplifier

Fig.2: the *Signal Amplifier* circuit is based around precision JFET-input op amp IC1. It uses a precision resistor and trimpot to provide a very accurate 10-times gain (+20dB) to boost the level of signals from devices such as arbitrary waveform generators. The 12V AC supply is boosted and regulated to 32V DC using a full-wave voltage-doubler configuration (D1 and D2), followed by a low-ripple adjustable linear regulator (REG1).

The underside of the LED body should be about 24mm above the top of the PCB. This will allow it to protrude through the box lid/front panel when the unit is assembled.

Your *Signal Amplifier* PCB is then virtually complete. The next step is to set the gain of its 10×/20dB range. Use a DMM with the best resistance accuracy possible.

Monitor the resistance between the junction of the 10kΩ resistor and 10μF capacitor near VR1, and the PCB's ground. Then adjust trimpot VR1 until this resistance is as close as possible to one-ninth the resistance of the 100kΩ 0.1% resistor.

If you're not confident of your DMM's accuracy, it may be easier to simply adjust the lower arm's resistance to measure 11,111Ω (11.111kΩ, or 100kΩ÷9). (Note that if you can measure both values on the same range, any proportional inaccuracy in the DMM itself should be cancelled out as it applies to both measurements.)

It's now time to test the completed PCB by connecting a source of 12V AC, such as an AC plugpack. LED1 should light up.

Measure the voltage between TP 32V and TP GND. You should get a reading close to 32V. If so, you can disconnect the power lead and put the PCB aside while you work on the box.

Parts list – Precision Signal Amplifier

1 double-sided PCB, code 04107191, 92 × 51mm
1 diecast aluminium box, 111 × 60 × 54mm [Jaycar HB5063]
1 12V AC plugpack (100mA or higher) with 2.1 or 2.5mm plug
1 SPST mini toggle switch (S1)
1 insulated BNC socket, single-hole panel mounting (CON1)
2 BNC sockets, single-hole panel mounting (CON2,CON3)
1 PCB-mount concentric DC socket, 2.1mm or 2.5mm inner diameter (to suit plugpack) (CON4)
4 25mm-long M3 tapped spacers
8 M3 × 6mm panhead machine screws
2 1mm PCB stakes (optional)
9 30mm lengths of hookup wire (to connect S1 and CON1-3 to the PCB)

Semiconductors
1 ADA4625-1ARDZ low-noise JFET-input op amp, SOIC-8 SMD package (IC1)
1 LM317M adjustable voltage regulator, SOT-223-3 SMD package (REG1)
4 1N5819 40V 1A schottky diodes (D1-D4)
1 3mm green LED (LED1)

Capacitors
2 470μF 25V RB electrolytic
1 47μF 35V RB low-ESR electrolytic
1 10μF 25V multi-layer ceramic (X5R 3216/1206 SMD)
2 1μF 100V polyester (radial leaded)
3 220nF 50V multi-layer ceramic (X5R 3216/1206 SMD)
1 100nF 50V multi-layer ceramic (X5R 3216/1206 SMD)
1 1.5pF 100V multi-layer ceramic (C0G 1206 or 0603 SMD)

Resistors (1% all SMD 3216/1206 SMD unless otherwise stated)
1 100kΩ 0.1% 0.25W axial leaded
1 100kΩ 3 10kΩ 1 5.6kΩ 1 3.3kΩ
1 820Ω 1 330Ω 1 240Ω 1 100Ω 2 51Ω
1 500Ω 15-turn horizontal trimpot (VR1)

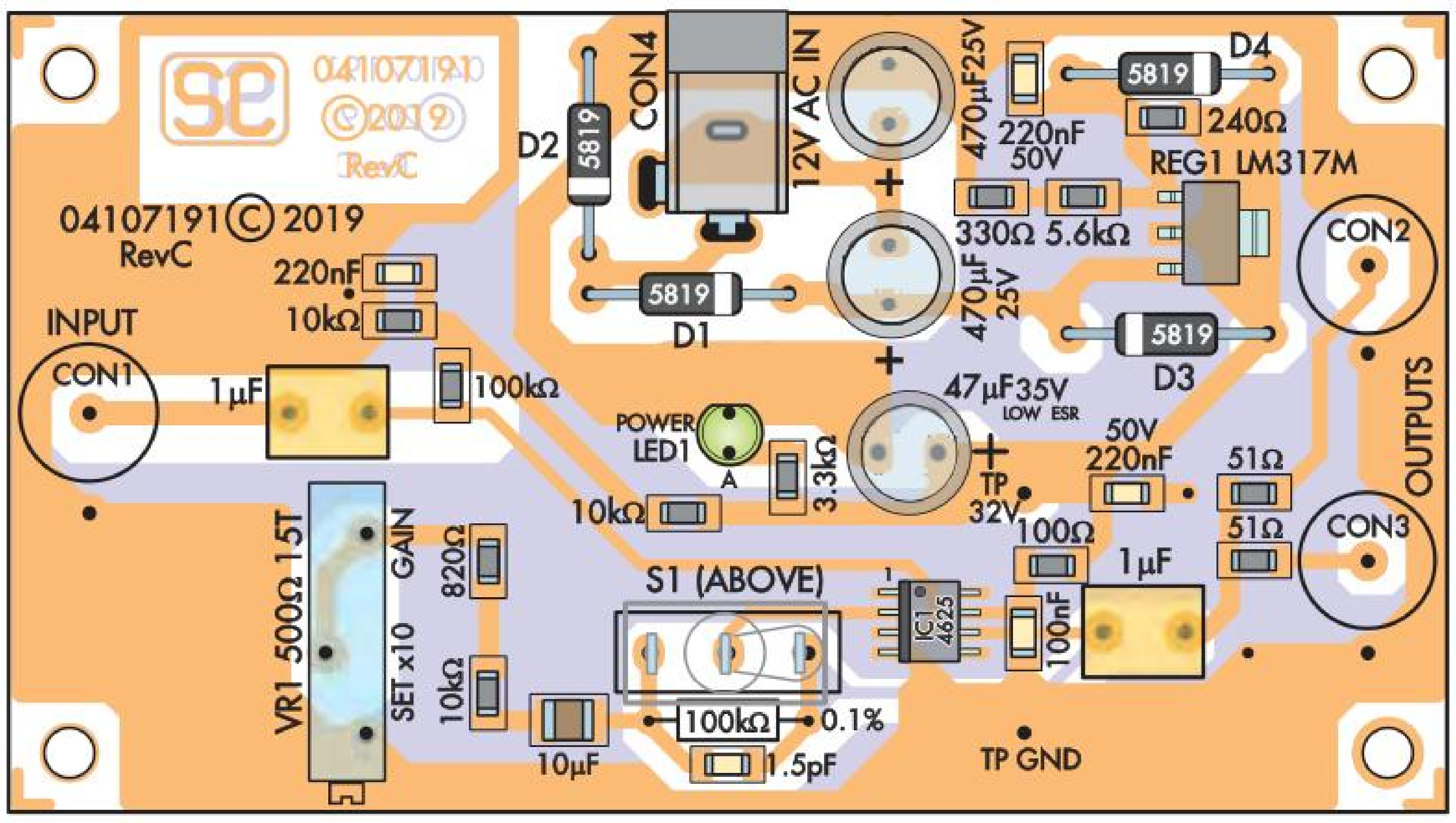
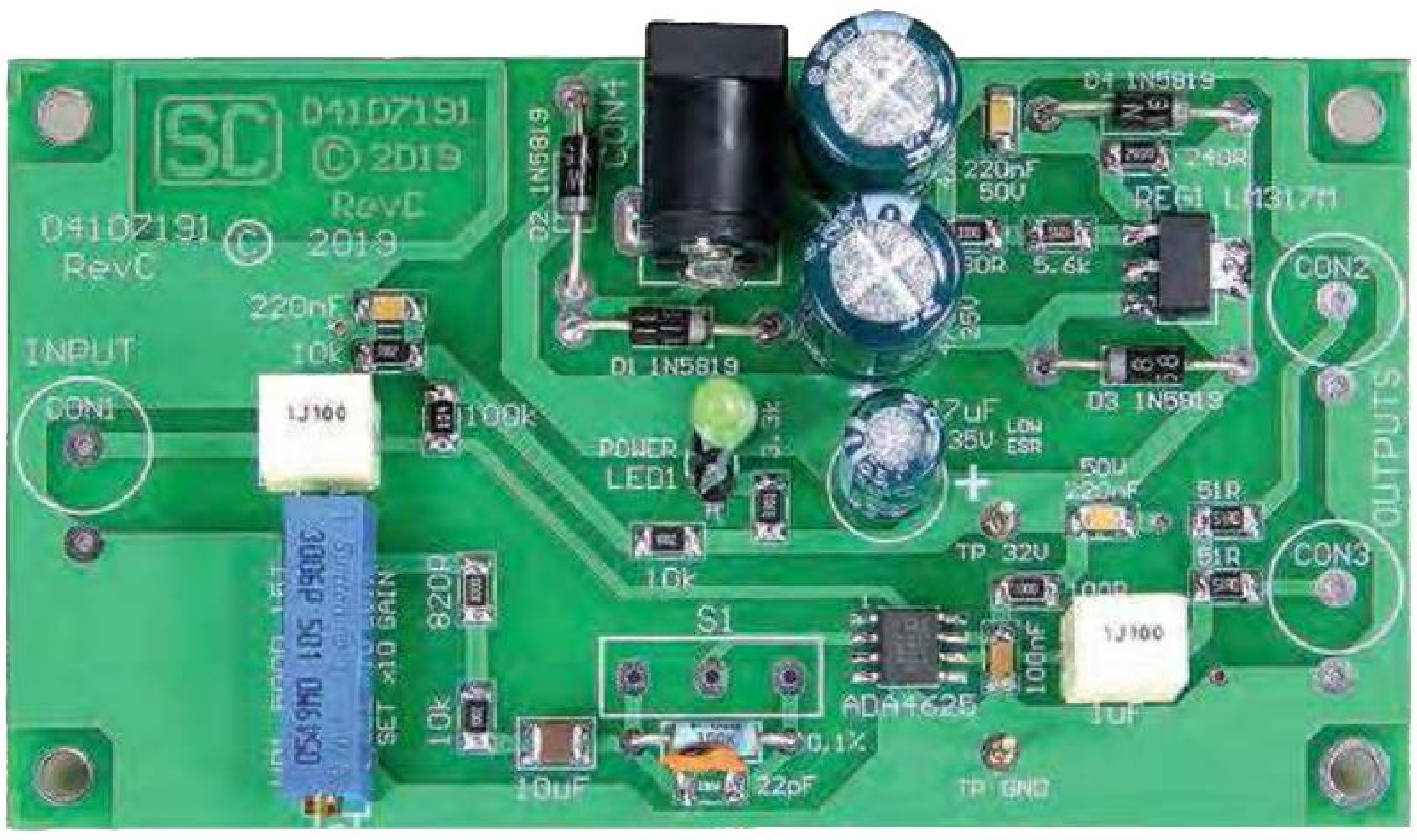

04107191 © 2019 RevC
INPUT
CON1
220nF
10kΩ
1µF
100kΩ
D2
5819
CON4
12V AC IN
470µF25V
220nF 50V
D4
5819
240Ω
REG1 LM317M
330Ω 5.6kΩ
470µF 25V
D1
5819
D3
5819
CON2
47µF35V LOW ESR
POWER LED1
A
3.3kΩ
50V 220nF
51Ω
51Ω
TP 32V
10kΩ
100Ω
CON3
OUTPUTS
VR1 500Ω 15T
GAIN
SET x10
820Ω
10kΩ
S1 (ABOVE)
IC1 4625
100nF
1µF
100kΩ
0.1%
10µF
1.5pF
TP GND

Fig.3: all of the components mount on this PCB, except for CON1-CON3 and switch S1. The design uses a mix of through-hole and surface-mounting parts. Fit them where shown here, being careful to ensure that IC1, LED1, diodes D1-D4 and the electrolytic capacitors are mounted with the correct polarity.



The almost-complete PCB just before final assembly. S1 and the connectors are not yet fitted because these mount on the front panel and connect to the PCB via short wire links. The PCB 'hangs' off the front panel via 25mm M3 tapped spacers, which are screwed to the four holes in the PCB corners.

Preparing the box

This is fairly straightforward. It involves drilling a total of nine holes in the box lid/front panel, another hole in the front of the box itself and then a larger hole (12mm diameter) in the box rear.

The locations and sizes of all these holes are shown in the cutting diagram, Fig.4.

After all of the holes have been drilled, cut and de-burred, you can attach a dress front panel to the lid, to give the *Signal Amplifier* a neat and user-friendly look. You can copy the front panel artwork shown in Fig.6, or download it as a PDF file from the October 2020 page of the *PE* website.

Then you can print it out and laminate it in a protective pouch to protect it from getting soiled.

It can then be attached to the box lid using double-sided adhesive tape. The final step is to use a sharp hobby knife to cut the holes in the dress panel, to match those in the lid underneath.

Now fit the three input and output BNC connectors to the front panel, along with switch S1, and then turn the panel over and solder short lengths of insulated hookup wire to the rear connection lugs of the connectors and S1.

Next, attach the four 25mm-long M3 tapped spacers to the corners of the front panel, using four 12mm-long M3 screws. Then you can cut the hookup wires soldered to CON1-CON3 and S1 to a length which will enable them to just pass through the PCB holes when the board is attached to the rear of the spacers.

Remove about 6mm of insulation from all of the wire ends, so that they can be easily soldered to the matching PCB pads.

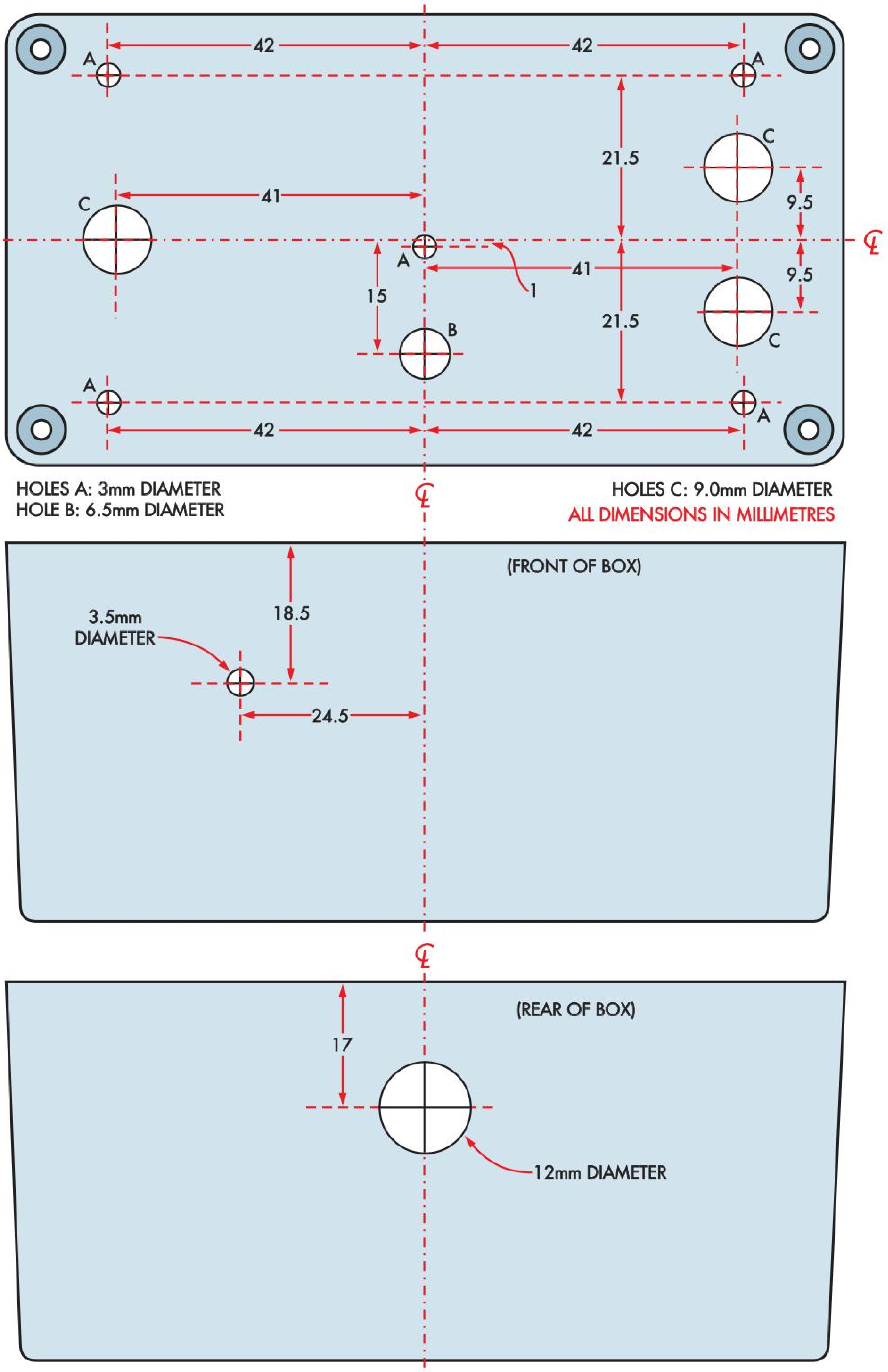
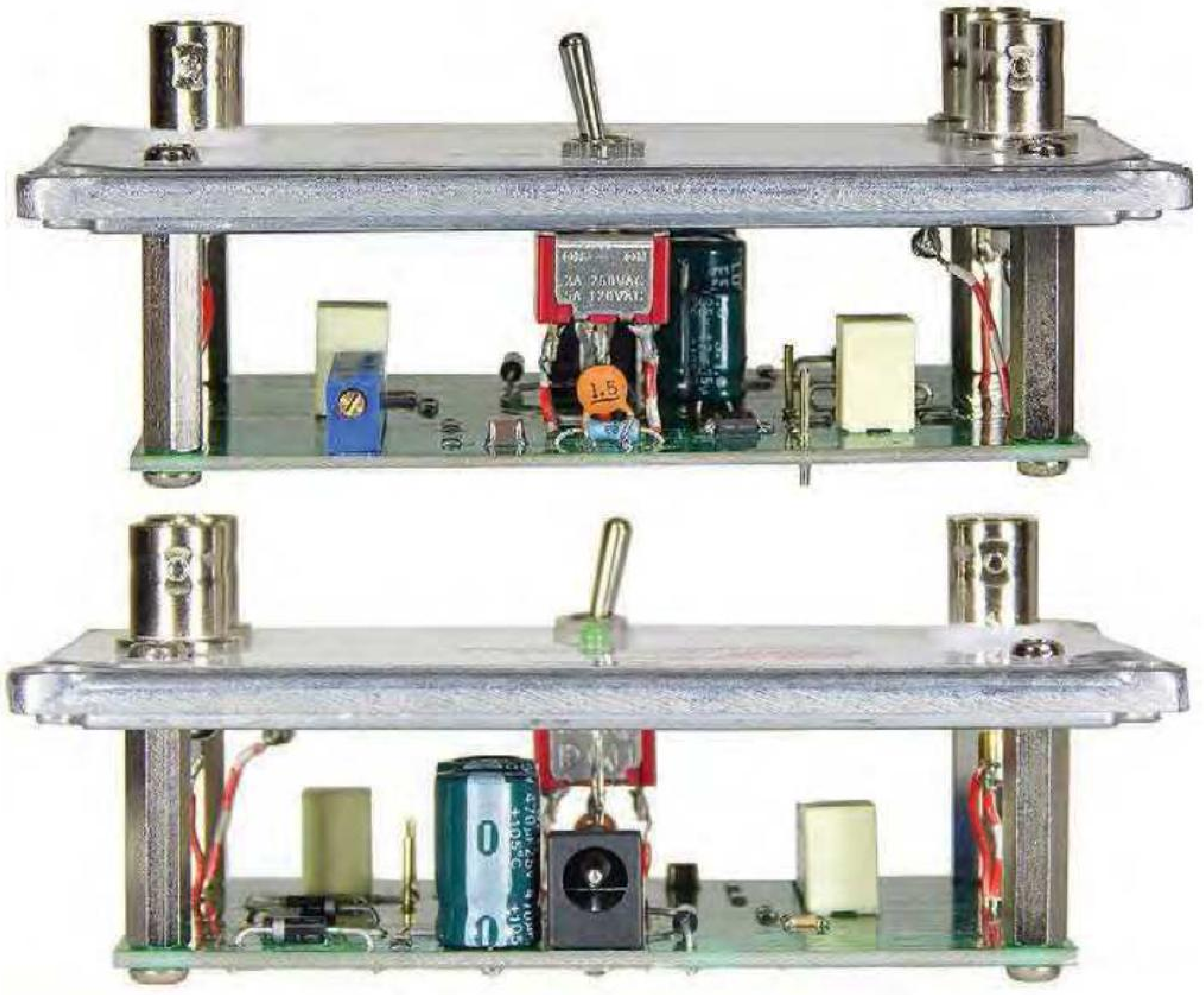

42
42
A
A
C
21.5
41
C
9.5
℄
A
41
1
9.5
15
21.5
B
C
A
A
42
42
HOLES A: 3mm DIAMETER
HOLE B: 6.5mm DIAMETER
℄
HOLES C: 9.0mm DIAMETER
ALL DIMENSIONS IN MILLIMETRES
(FRONT OF BOX)
18.5
3.5mm DIAMETER
24.5
℄
(REAR OF BOX)
17
12mm DIAMETER

Fig.4: here are the locations and sizes of the holes that need to be drilled in the diecast aluminium enclosure. For the larger holes, it's best to start with a smaller pilot hole (eg, 3mm) and then enlarge it to size using either a stepped drill bit, a series of larger drills or a tapered reamer. That ensures accurate positioning and a clean, round hole. You can copy this diagram and attach it to the box using tape to use it as a template. (It can also be downloaded from the October 2020 page of the *PE* website.)



Above are two views of the assembled unit from the front (top) and the rear (bottom).

End-on view from the input end. The ×10 gain is calibrated via the multi-turn pot (blue component) in the foreground.

After bending these wires so that their ends are positioned to meet with the holes in the PCB, offer up the PCB assembly to the spacers on the rear of the panel.

With a bit of jiggling, you should be able to get all of the wires to pass through their matching holes. You can then attach the PCB to the spacers using four more 6mm-long M3 screws, up-end the assembly and solder each of the wires to its PCB pad.

Your *Signal Amplifie*r is now complete, and should look like the one shown in our photos.

All that remains is to lower the lid-and-PCB assembly into the box and fasten them together using the four M4 countersunk-head screws supplied with it.

Checkout and use

At this stage, your *Signal Amplifier* should be ready for use. Remember that it can deliver a maximum output voltage of 30V peak to peak or 10.6V RMS, assuming that it is feeding a high-impedance load, of 50kΩ or more.

If the load impedance is much lower, the maximum output amplitude will be slightly reduced.

Note that you can check and adjust its calibration even after it has been sealed in its box. To do this, you will need an audio oscillator or function generator and a DMM with a true-RMS AC voltage range with reasonable accuracy and resolution.

Set your oscillator or function generator to produce a sinewave at 1kHz and around 1V RMS, then connect its output to the input of the *Signal Amplifier*. Then connect your DMM's input to one of the *Signal Amplifier* outputs.

With the *Signal Amplifier's* gain set to unity (A=1), power it up and your DMM should indicate an AC voltage very close to 1.00V. If not, you may need to tweak the output of your oscillator/generator until this reading is achieved.

Then all you have to do is select the *Signal Amplifier's* ×10 range, at which point the DMM reading should jump to 10.000V – or very close to it. Now adjust trimpot VR1 with a small screwdriver or alignment tool (through the small hole in the front of the box), until the DMM is reading exactly 10.000V.

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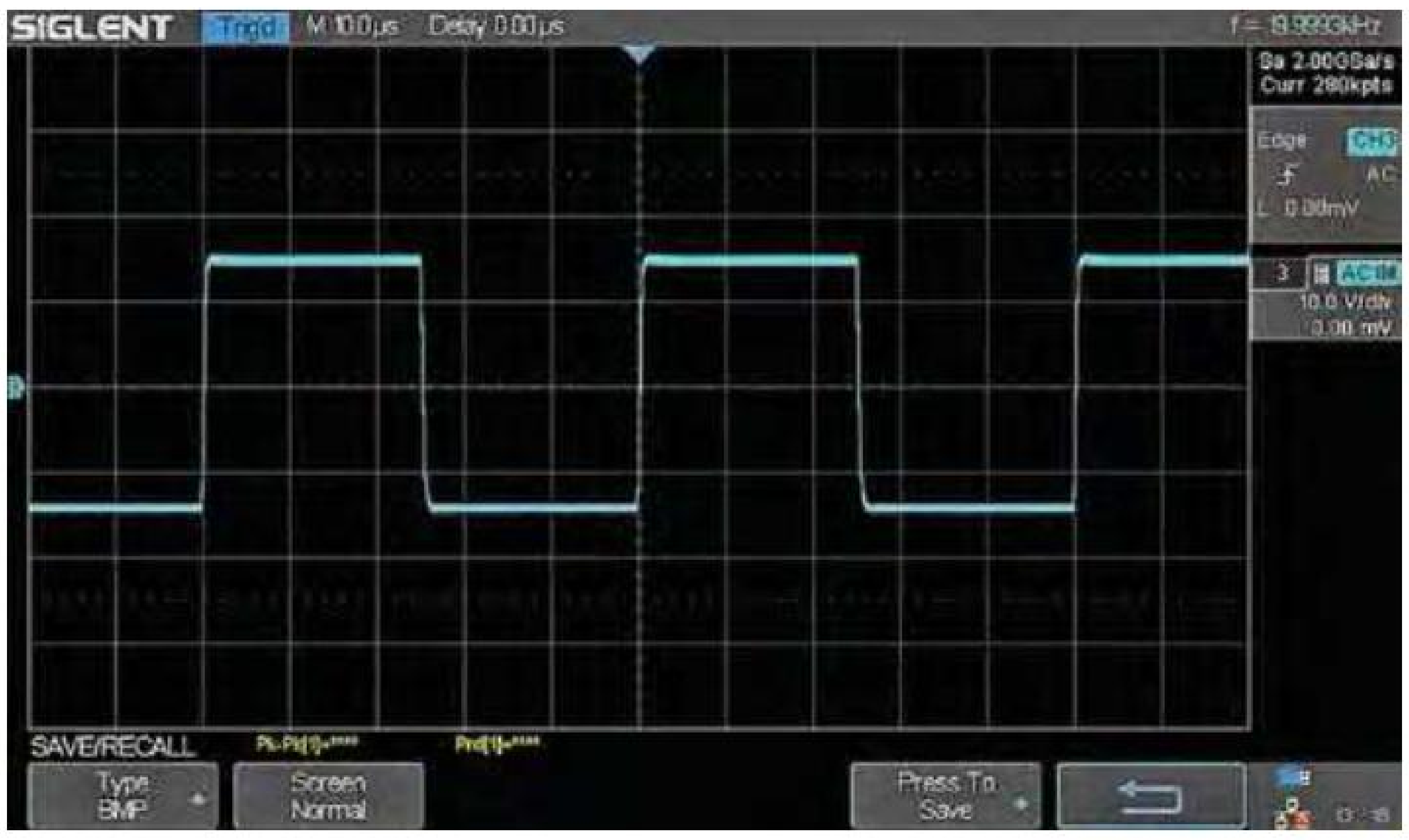
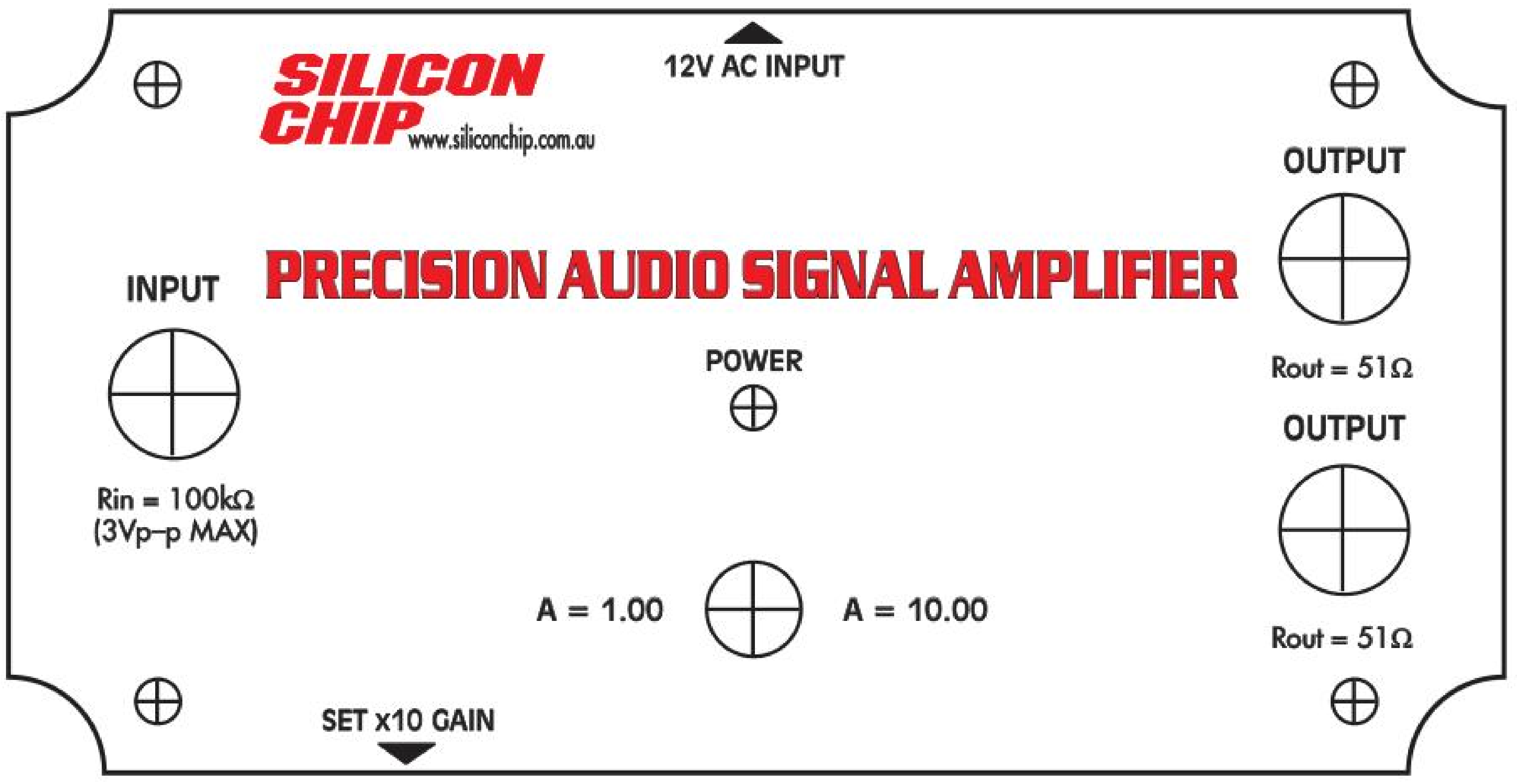

SIGLENT
Trig'd
M 10.0μs
Delay 0.00μs
f= 19.9993kHz
Sa 2.000GSa/s
Curr 280kpts
Edge
CH3
AC
L 0.00mV
AC IM
10.0 V/div
3.00 mV
SAVE/RECALL
Type BMP
Screen Normal
Press To Save

Fig.5: this scope grab of the unit's output with a full-swing (32V peak-to-peak) 20kHz square wave demonstrates the fast slew rate and quick settling time of the ADA4625 op amp.

You can see that there is minimal rounding and overshoot after each transition, and it settles close to the target value in well under 1μs.


SILICON CHIP
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12V AC INPUT
OUTPUT
PRECISION AUDIO SIGNAL AMPLIFIER
INPUT
POWER
Rout = 51Ω
OUTPUT
Rin = 100kΩ
(3Vp–p MAX)
A = 1.00
A = 10.00
Rout = 51Ω
SET x10 GAIN

Fig.6: this 1:1 front panel artwork can be copied and fixed to the lid or it can be downloaded from the October 2020 page of the *PE* website, printed and then applied.

ARDUINO-BASED DIGITAL AUDIO MILLIVOLTMETER

PC (USB) & 5V DC INPUT
SAMPLING
INTERCEPT ADJUST
0.038 mV (U/B)
= -88.47 dBV (L)
CALIBRATE HI RANGE
SILICON CHIP
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HIGH RESOLUTION AUDIO MILLIVOLTMETER
PUSH
INPUT SELECT
BALANCED
UNBALANCED
UNBAL INPUT (Zin > 10kΩ)
BAL INPUT (Zin = 760kΩ)

ARDUINO

by Jim Rowe

Low cost, easy to build, highly accurate: an essential piece of test equipment!

If you're involved in audio – at any level – you really must have an audio millivoltmeter in your test gear arsenal. Once you've used one, you'll wonder how you ever managed without it. It's useful for setting up and calibrating audio systems, doing performance measurements and troubleshooting audio equipment, and much more. This one doesn't just measure low-level signals. It provides high-resolution measurements of balanced or unbalanced audio signals from below –85dBV (56µV RMS) to above +35dBV (60V RMS)! It's easy to build and has automatic range switching and can log data to a PC.

We decided to design a new *Audio Millivoltmeter* because we wanted one which worked over a very wide range of signal amplitudes with excellent accuracy and resolution.

We also wanted to have the ability to measure balanced or unbalanced audio signals without the need for any additional hardware. But, most of all, we wanted it to be easy to build and would fit in a compact case.

So why build this one instead of our previous audio millivoltmeter (March 2011 – *Low-Cost Digital Audio Millivoltmeter*)? Well, for many reasons this new unit makes that old one obsolete:

- It can measure smaller signals and much larger signals
- It has much better resolution
- Its frequency response (on both ranges) is much better (see Fig.1)
- It has a built-in balanced input (no separate converter required)
- It does not require manual range selection
- It runs off USB power
- It is quite a bit smaller.

Some of the improvements in this version are due to our use of an Arduino Nano MCU module for control, while most of the performance improvements are due to our use of an LTC2400 24-bit analogue-to-digital converter (ADC).

Features and specifications

• Description	*A compact high-resolution digital audio millivolt/voltmeter with balanced and unbalanced inputs, backlit LCD readout, automatic range switching and the ability to send its data to a PC.*
• Unbalanced measurement range	*from <56µV RMS (–85dBV) to 60V RMS (+35dBV)*
• Balanced measurement range	*from <56µV RMS (–85dBV) to 600mV RMS (–4.5dBV)*
• Frequency range	*5Hz-110kHz (+0/–3dB); 20Hz-70kHz (+0/–0.5dB); 50Hz-45kHz (+0/–0.1dB)*
• Resolution	*24 bits (1 part in 16,777,215)*
• Measurement linearity	*±0.3dB*
• Basic accuracy	*approximately ±0.1% after calibration*
• Input impedance	*1MΩ/10kΩ (unbalanced input) or 760kΩ (balanced input)*
• Maximum input level	*as per measurement ranges*
• Power supply	*5V DC via USB mini Type-B socket, either from a USB charger or a PC USB port*
• Current drain	*<78mA (390mW at 5V)*

This gives much higher measurement resolution than the 10-bit ADC built into most Arduinos.

The result is a unit that's much more convenient to use, with higher performance and it fits into a diecast box measuring only 119 × 94 × 57mm. That's less than half the volume of the earlier version.

We estimate the total cost for everything you'll need to build this project to be under £125. That compares more than favourably with what you'd pay for a similar commercial instrument.

To give you an idea of why you might want to measure down to –85dBV, if you have a 100W amplifier which can drive 8Ω loads, at full power then it's delivering 28.28V RMS ($\sqrt{100W \times 8\Omega}$) across the speaker. That equates to +29dBV.

For such an amplifier, a noise level of –85dBV would therefore mean a signal-to-noise ratio of 114dB (85dB + 29dB). A good amplifier can do that.

So if you had a meter which couldn't measure down to –85dB, you couldn't come close to getting an accurate measurement of the signal-to-noise ratio of such an amplifier.

The best our 2011 design could achieve was –76dBV, limiting you to SNR measurements of no better than about 105dB for a 100W amp, and considerably worse than that for lower-powered amplifiers, or line-level devices.

How it works

Fig.2 is a simplified block diagram of the new meter. At its heart is IC3, an Analog Devices AD8307 logarithmic amplifier/detector. This is the same device used in our earlier meter.

The AD8307 has impressive specifications: it can convert AC signals into a DC voltage equivalent, following a logarithmic 'law' of 25mV per dB (typically linear to within ±0.3dB) and with a span of just on 100dB. The device also operates up to around 500MHz, so it's just 'idling' at audio frequencies.

In the new meter, we are feeding IC3's output to IC4, an LTC2400 24-bit delta/sigma ADC. This measures the output of IC3 relative to an accurate 2.500V DC provided by an LT1019 bandgap voltage reference. The resulting 24-bit digital samples are passed to the Arduino Nano via SPI (serial peripheral interface).

The microcontroller then processes the samples to calculate the corresponding measurements, which are displayed on the LCD module shown at upper right in Fig.2.

They're also sent out via the D– and D+ lines of the USB socket at lower right, for logging via a PC if required. The micro indicates when sampling is taking place by lighting LED1.

The elements on the left-hand side of Fig.2 have been added to provide input buffering, low-pass filtering (to reject RF or other unwanted signals), range selection and selection between the unbalanced and balanced inputs.

IC1 is an AD629B high-common-mode-voltage-rejecting difference amplifier, used to convert the balanced input signals from XLR socket CON1 into an unbalanced signal. Switch S1a then selects between the unbalanced signals from either CON2 or the output of IC1, with the other half of the double-pole switch (S1b) allowing the micro to detect which input is currently selected.

The signal then goes into the range switching section, where a reed relay controlled by the micro via transistor

Some potential uses

- ☑ **Audio performance measurements (signal-to-noise ratio, frequency response, sensitivity, power output, channel separation, crosstalk, amplifier gain)**
- ☑ **Crossover adjustment**
- ☑ **Equalisation and room response adjustments (in combination with a microphone and preamp)**
- ☑ **Amplifier calibration**
- ☑ **Amplifier and preamplifier troubleshooting and repair**

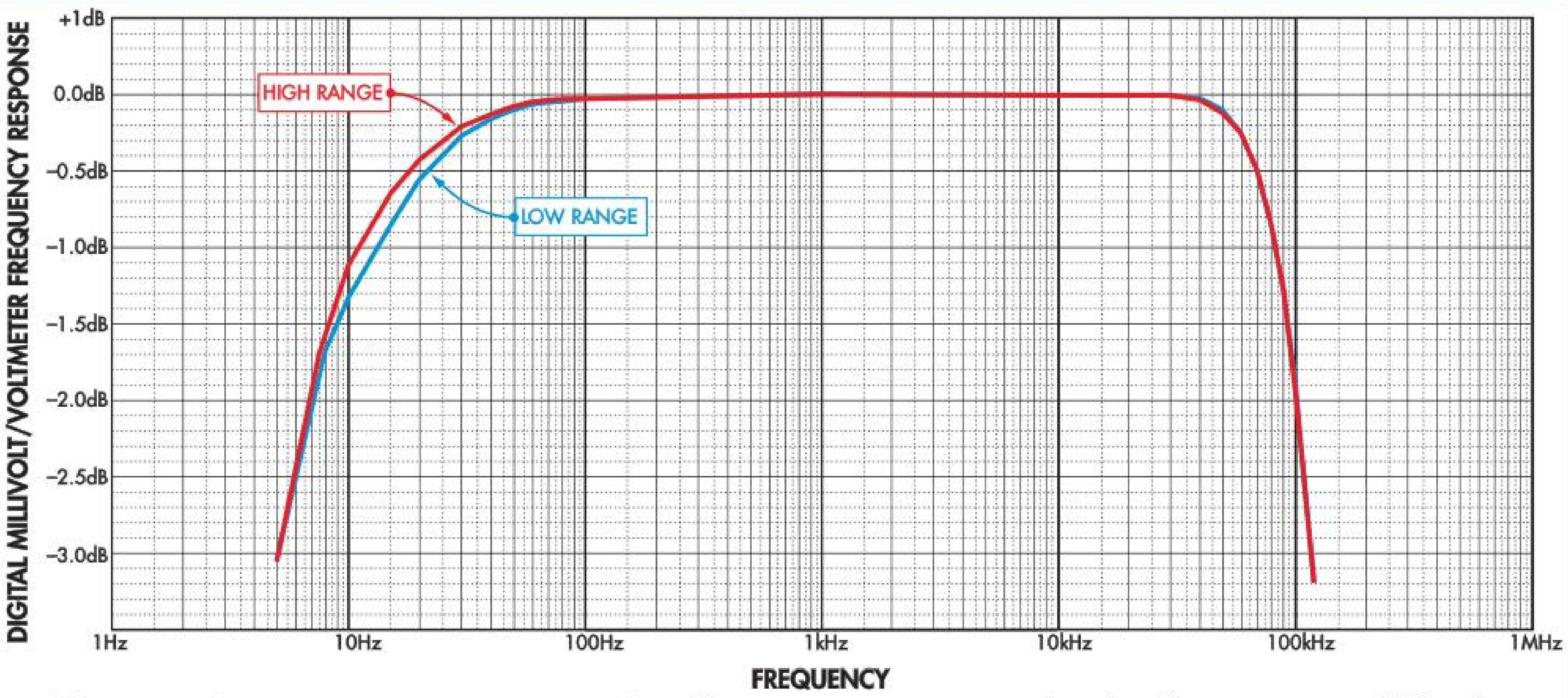

Fig.1: a frequency response plot for our prototype in the low range (blue) (measured at 600mV RMS) and high range (red). This demonstrates that the reading is within 0.5dB of the actual signal amplitude over the entire audible range and beyond. It's within 0.1dB from 50Hz to 45kHz.

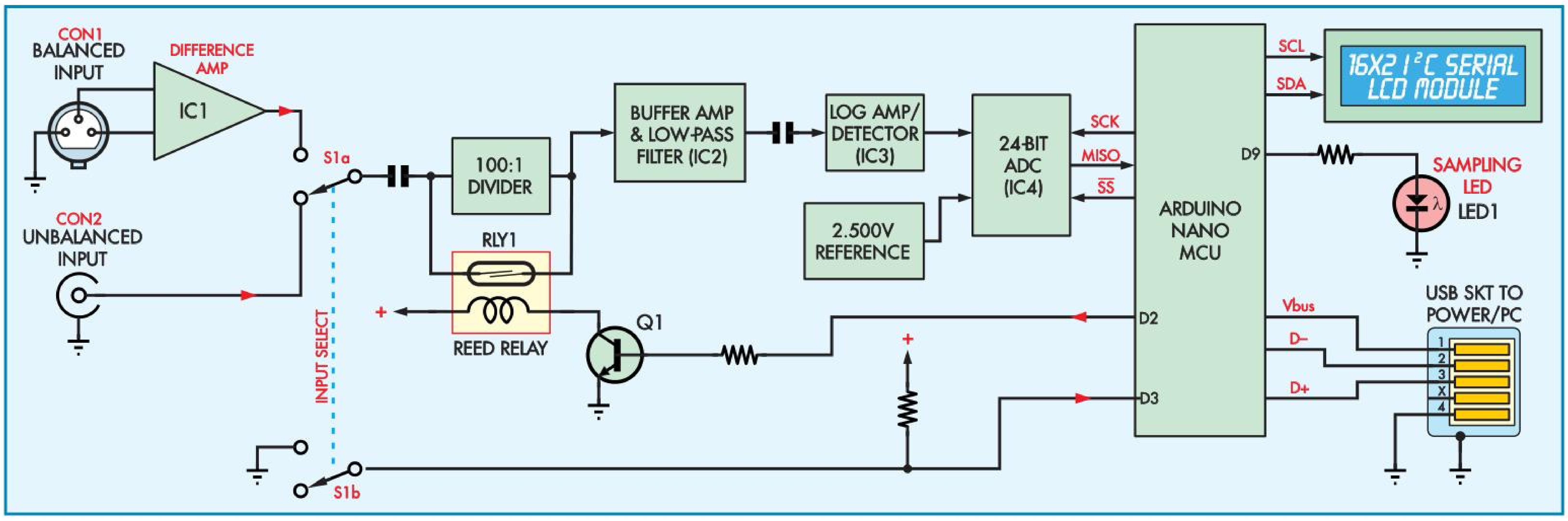
CON1
BALANCED
INPUT
DIFFERENCE
AMP
IC1
CON2
UNBALANCED
INPUT
S1a
INPUT SELECT
S1b
100:1
DIVIDER
RLY1
REED RELAY
Q1
BUFFER AMP
& LOW-PASS
FILTER (IC2)
LOG AMP/
DETECTOR
(IC3)
24-BIT
ADC
(IC4)
2.500V
REFERENCE
SCK
MISO
SS
ARDUINO
NANO
MCU
SCL
SDA
D9
D2
D3
16X2 I²C SERIAL
LCD MODULE
SAMPLING
LED
LED1
Vbus
D–
D+
USB SKT TO
POWER/PC
1
2
3
X
4

Fig.2: this block diagram shows the operating principle of the *Meter*. IC1 converts a balanced signal to unbalanced and S1 selects between the two inputs. The signal then either passes through RLY1 or a 100:1 divider, depending on whether Q1 (and therefore RLY1) is energised, giving the unit its two ranges. The signal is then buffered, filtered and fed to the logarithmic detector before passing to the ADC and onto the Arduino.

Q1 is used to select between either the input signal divided by 100 (for the high range, up to 60V), or bypassing the divider (for the low range).

The signal is then fed to IC2, a dual op amp with the first stage used as a unity-gain buffer and the second stage as a low-pass filter.

This removes, or at least significantly reduces, any noise (including digital switching artefacts from the control circuitry) which may be induced into the analogue signal.

The full circuit

You'll find more details in the main circuit diagram (Fig.3). The signal from the balanced input at CON1 is filtered using a common-mode choke (T1) and a 47pF capacitor to remove RF signals, before being coupled via two high-voltage capacitors to the inputs of IC1, the balanced-to-unbalanced converter. This allows for balanced common-mode signals up to 400V peak from earth.

A 2.5V bias signal is applied to the REF– and REF+ inputs of IC1 (pins 1 and 5), biasing its input signals to half of the 5V supply, to allow for a symmetrical signal swing before it runs into clipping.

The signal from the unbalanced input (CON2) is also RF filtered using inductor L1 and a 22Ω series resistor and 22pF capacitor to ground.

The output from selector switch S1 is AC-coupled to the precision 100:1 voltage divider, the upper portion of which is shorted out when the contacts of RLY1 are closed for measuring lower

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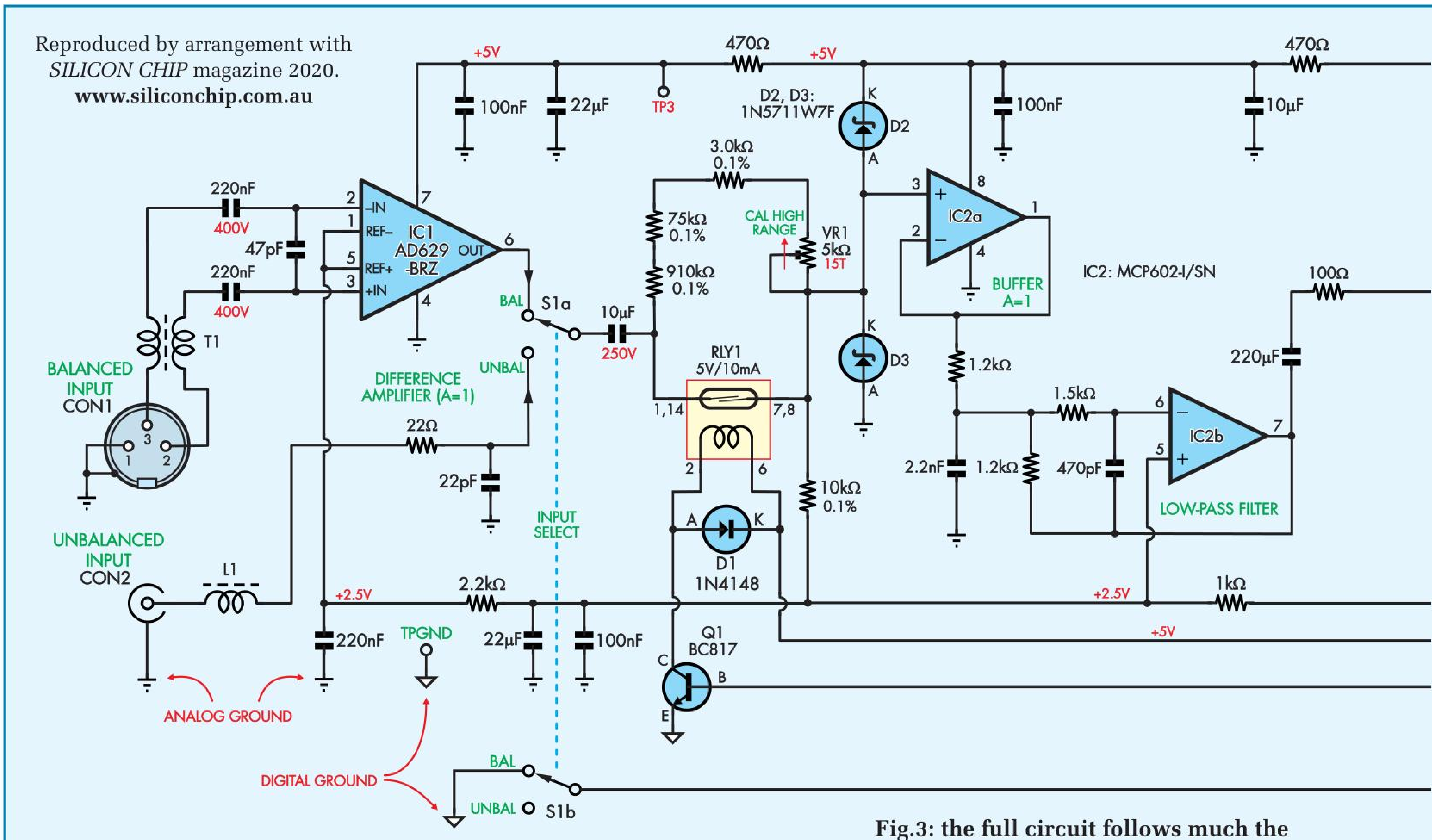

+5V
100nF
22µF
TP3
470Ω
+5V
D2, D3:
1N5711W7F
K
D2
A
100nF
470Ω
10µF
220nF
400V
47pF
220nF
400V
2
1
5
3
–IN
REF–
REF+
+IN
IC1
AD629
-BRZ
OUT
7
4
6
T1
BALANCED
INPUT
CON1
3
1
2
BAL
S1a
UNBAL
DIFFERENCE
AMPLIFIER (A=1)
22Ω
22pF
10µF
250V
3.0kΩ
0.1%
75kΩ
0.1%
910kΩ
0.1%
CAL HIGH
RANGE
VR1
5kΩ
15T
RLY1
5V/10mA
1,14
7,8
2
6
10kΩ
0.1%
A
K
D1
1N4148
K
D3
A
3
2
8
4
1
IC2a
BUFFER
A=1
IC2: MCP602-I/SN
1.2kΩ
1.5kΩ
6
5
7
IC2b
2.2nF
1.2kΩ
470pF
LOW-PASS FILTER
220µF
100Ω
+2.5V
1kΩ
+5V
INPUT
SELECT
UNBALANCED
INPUT
CON2
L1
+2.5V
2.2kΩ
220nF
TPGND
22µF
100nF
Q1
BC817
C
B
E
ANALOG GROUND
DIGITAL GROUND
BAL
UNBAL
S1b

Digital Audio Millivolt/Voltmeter

Fig.3: the full circuit follows much the same pattern as Fig.2, but you can see that some details were left out of the earlier diagram, such as the input RF filtering. VR1 allows the 100:1 divider to be accurately trimmed, while VR2

level signals. Trimpot VR1 is used to 'fine-tune' the divider for calibrating the *Meter's* HIGH range.

The way the divider works, and the reason for selecting these exact component values, is shown in more detail in Fig.4.

The values are selected so that trimpot VR1 can be used to set the divider ratio to precisely 100:1 without restricting its rotation to a narrow portion of its range. VR1 can compensate for within-tolerance variations in the four 0.1% tolerance fixed resistors.

Note that as well as forming the lower leg of the divider for the *Meter's* HIGH range, the 10kΩ 0.1% resistor also forms the input resistance for the *Meter's* LOW range, for the unbalanced input.

That's because when RLY1 is switched on to short out the divider's upper arm for the LOW range, the lower part of the divider still provides the DC bias for input pin 3 of IC2a.

Pin 21 of the Arduino (the D3 digital input) is used to monitor the position of S1, while pin 20 (digital output D2) controls the range selection relay (RLY1) via NPN transistor Q1. Diode D1 protects transistor Q1 from damage due to the back-EMF generated by the coil of RLY1 when it switches off.

Schottky diodes D2 and D3 protect IC2a from overload damage, by clamping its pin 3 input voltage within a few hundred millivolts of the supply rails, even if the input signal amplitude is too high for the *Meter* to measure accurately.

The purpose of IC2a is to buffer the signal from the divider to provide a low-impedance source for the following low-pass filter, which is built around the other half of the dual op amp, IC2b.

This is a second-order (–12dB/octave) 'multiple feedback' low-pass filter with a –3dB point of around 52kHz. This was chosen to give a very flat response up to 20kHz, then a steep roll-off above audio frequencies.

This filter is important since, as stated earlier, the log converter (IC3) has a wide bandwidth of up to 500MHz. So any digital noise or RF picked up before this point will add to the signal being detected and give erroneous readings. Therefore, we want to ensure that all ultrasonic frequency signals are severely attenuated.

This filter type and its values were chosen carefully for this role, as a multiple-feedback filter has a significant advantage over the more common Sallen-Key type in that it still provides excellent attenuation for signals above the op amp's bandwidth, and it is far less reliant on said bandwidth to provide the expected filter attenuation.

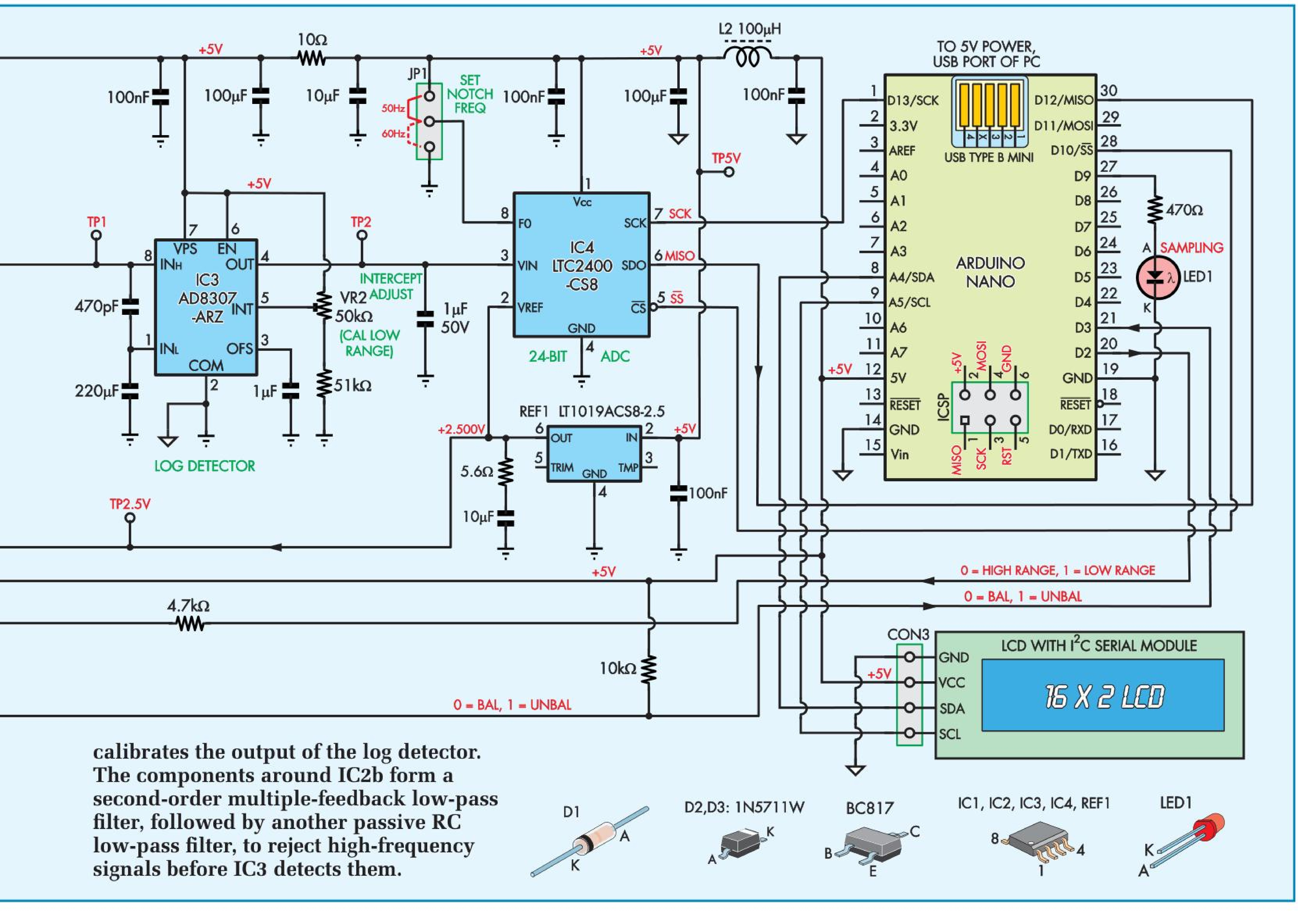
A second-order multiple-feedback resistor needs just one more resistor than a Sallen-Key type, which is well worth it for its superior high-frequency attenuation.

The inputs of IC2a and IC2b are biased to the 2.5V rail, both through its connection to the bottom of the switchable voltage divider ladder, as well as it being fed directly to pin 5 of IC2b.

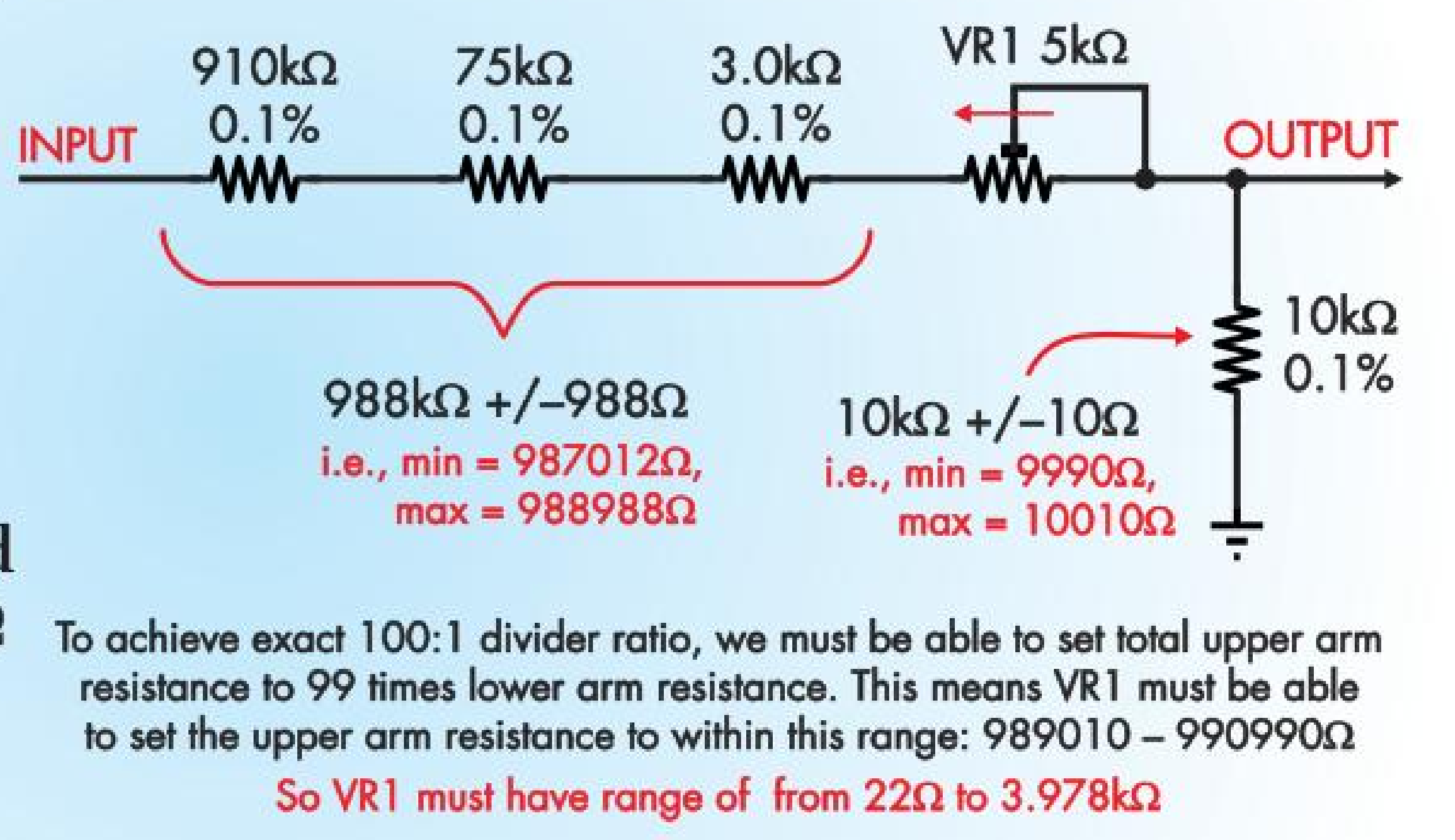
Again, this biases the AC signal fed to these rail-to-rail op amps so that it swings symmetrically within the 5V supply.

The audio signal is then AC-coupled to input pin 8 of the AD8307 log detector. A 100Ω series resistor provides additional RF filtering, in combination with the 470pF capacitor between its pins 8 and 1. Pin 1 is grounded via a 220µF capacitor, as we are not feeding differential signals to this chip. The INL input sits at the chip's DC bias level while the INH input swings above and below that voltage.

Trimpot VR2 allows us to adjust IC3's 'intercept' point, calibrating the *Meter's* LOW measurement range. A 1µF capacitor smoothes the logarithmic output voltage from pin 4, and


+5V
10Ω
100nF
100µF
10µF
JP1
SET NOTCH FREQ
50Hz
60Hz
L2 100µH
TO 5V POWER, USB PORT OF PC
TP5V
+5V
TP1
TP2
IC3 AD8307 -ARZ
VPS
EN
INH
OUT
INT
INL
OFS
COM
470pF
220µF
1µF
51kΩ
VR2 50kΩ (CAL LOW RANGE)
INTERCEPT ADJUST
LOG DETECTOR
1µF 50V
IC4 LTC2400 -CS8
Vcc
F0
SCK
VIN
SDO
VREF
CS
GND
SCK
MISO
SS
24-BIT
ADC
REF1 LT1019ACS8-2.5
OUT
IN
TRIM
TMP
GND
+2.500V
5.6Ω
10µF
100nF
TP2.5V
+5V
4.7kΩ
10kΩ
0 = BAL, 1 = UNBAL
0 = HIGH RANGE, 1 = LOW RANGE
0 = BAL, 1 = UNBAL
ARDUINO NANO
USB TYPE B MINI
D13/SCK
3.3V
AREF
A0
A1
A2
A3
A4/SDA
A5/SCL
A6
A7
5V
RESET
GND
Vin
D12/MISO
D11/MOSI
D10/SS
D9
D8
D7
D6
D5
D4
D3
D2
GND
RESET
D0/RXD
D1/TXD
ICSP
MISO
SCK
RST
MOSI
GND
+5V
470Ω
SAMPLING
LED1
A
K
CON3
GND
VCC
SDA
SCL
LCD WITH I²C SERIAL MODULE
16 X 2 LCD
D1
D2,D3: 1N5711W
BC817
IC1, IC2, IC3, IC4, REF1
LED1
A
K
B
C
E
8
4
1

calibrates the output of the log detector. The components around IC2b form a second-order multiple-feedback low-pass filter, followed by another passive RC low-pass filter, to reject high-frequency signals before IC3 detects them.



To achieve exact 100:1 divider ratio, we must be able to set total upper arm resistance to 99 times lower arm resistance. This means VR1 must be able to set the upper arm resistance to within this range: 989010 – 990990Ω

So VR1 must have range of from 22Ω to 3.978kΩ

Fig.4: the details of the precision 100:1 divider. Starting with the choice of a 10kΩ 0.1% resistor in the bottom leg (which can have a value from 9.99kΩ to 10.01kΩ), that means we need a total resistance in the upper leg of 990kΩ±990Ω. Taking into account the tolerance of the fixed resistors in that upper leg, a 5kΩ potentiometer gives sufficient scope for adjusting for precisely the right attenuation factor.

this is then fed to the analogue input of IC4, the 24-bit ADC.

JP1, connected to pin 8 of IC4, changes the ADC's internal sampling frequency to provide a 'notch' for rejecting either 50Hz or 60Hz 'hum' in the signal from IC3.

So for UK use it would be set in the upper (50Hz) position, while in the US and other countries with 60Hz mains power, you'd set it in the lower position.

REF1 provides a very stable 2.5V reference to IC4, necessary for it to operate with the high precision possible for a 24-bit ADC. This means its resolution is 149nV (2.5V ÷ 2^{24}), so the limiting factor in its performance will be system noise.

The reference has an initial tolerance of ±0.05%, which equates to ±1.25mV. REF1's output also provides the 2.5V biasing for IC1 and IC2 mentioned earlier.

The reference output is stabilised by a Zobel network (5.6Ω and 10µF), as recommended in its data sheet.

The Arduino Nano communicates with the ADC (IC4) with the standard SPI pins (ie, pins D10, D12 and D13) while communication with the LCD is via an I^2C bus at pins A4/SDA and A5/SCL. Sampling LED1 is driven from the D9 digital output.

Construction

Most of the circuitry and components of the new *Meter* (including the Arduino Nano) are mounted on a PCB measuring 109 × 84mm, and which is coded 04106191.

The only components not mounted on the PCB are the LCD module, the input connectors and input selector switch S1. These mount on the box front panel and connect to the PCB via short lengths of wire.

Some of the components on the PCB are of the through-hole variety and somewhat larger than the SMD components. So it's best to fit the smaller SMD parts first.

The location and orientation of all parts are shown on the PCB overlay diagram (Fig.5), but you can also refer to the photos. Note though that there may be some slight differences between the prototype and final PCBs.

There are no fine-pitch SMD parts; all of them are reasonably generous in terms of size and pin spacings, so they are not difficult to handle. Start by fitting all the SMD passives (resistors and capacitors), except for those which are right next to one of the SMD ICs, as these would otherwise make fitting the latter more tricky.

The usual technique is to tack one side of the component onto its pad, make sure it is sitting flat on the board and properly aligned, then solder the opposite pad (after waiting long enough for the first joint to solidify). Then wait a little longer and refresh the first joint with a little extra solder or flux paste.

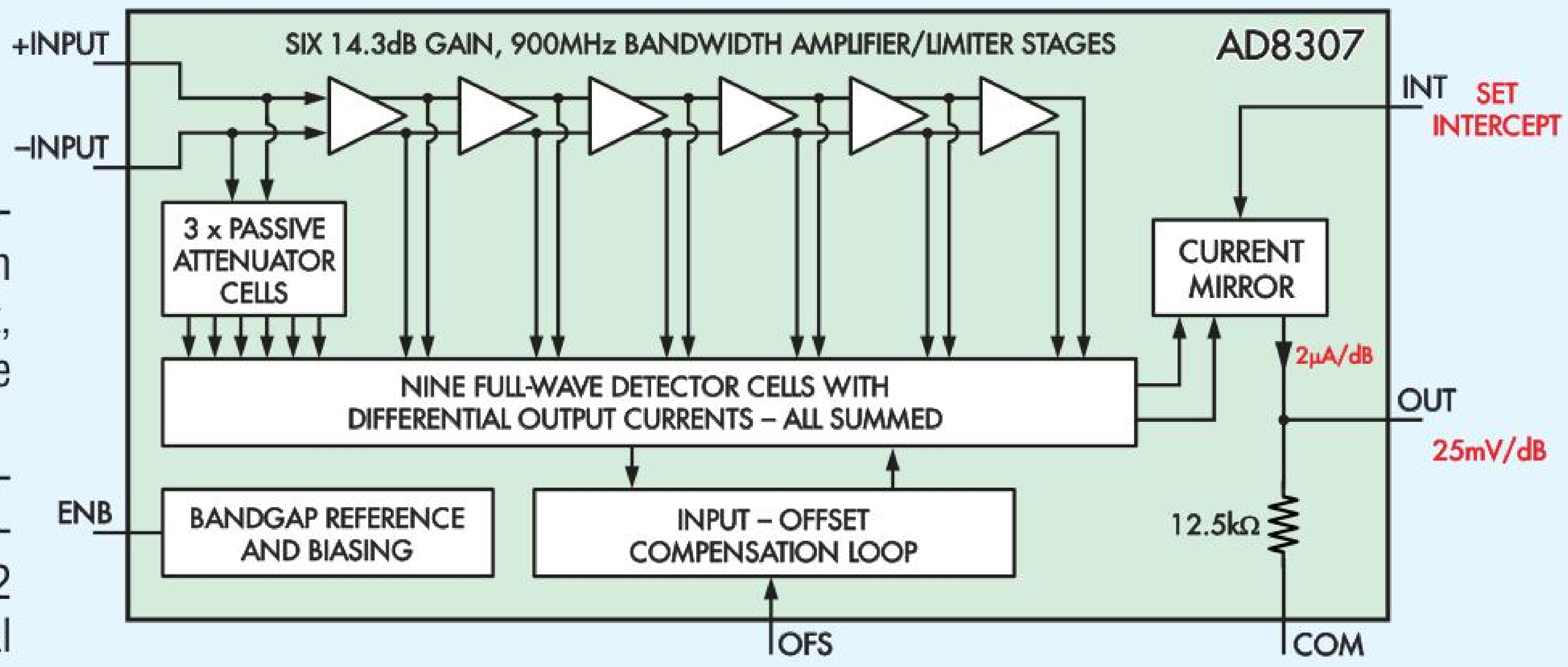
With those passives all in place, you can install the five SMD ICs. In each case, they must be oriented correctly, so find the pin 1 dot or divot on the top face, and make sure it's facing as shown in Fig.5. If you can't find the dot, pin 1 is normally also indicated by a chamfered edge on just that side of the IC.

Again, locate the IC and tack one pin down before soldering the other seven pins, then refresh that initial joint. The pins are spaced far enough apart to be soldered individually. If you accidentally form a solder bridge between two pins, add a little flux paste and then clean it up using solder wick.

The AD8307 logarithmic amplifier/detector

Logarithmic amplifier/detector ICs are a fairly specialised but quite useful device. You can get an idea of how they work from the diagram at right, which gives a simplified view of what's inside the AD8307 device.

The incoming AC signals pass through six cascaded wideband differential amplifier/limiter stages, each of which has a gain of 14.3dB (about 5.2 times) before it enters limiting. This gives a total gain of about 86dB, or around 20,000 times.


+INPUT
–INPUT
SIX 14.3dB GAIN, 900MHz BANDWIDTH AMPLIFIER/LIMITER STAGES
AD8307
INT
SET INTERCEPT
3 x PASSIVE ATTENUATOR CELLS
CURRENT MIRROR
2µA/dB
NINE FULL-WAVE DETECTOR CELLS WITH DIFFERENTIAL OUTPUT CURRENTS – ALL SUMMED
OUT
25mV/dB
ENB
BANDGAP REFERENCE AND BIASING
INPUT – OFFSET COMPENSATION LOOP
12.5kΩ
OFS
COM

The outputs of each amplifier/limiter stage are fed to a series of nine full-wave detector cells, along with similar outputs from three cascaded passive 14.3dB attenuator cells connected to the input of the first amplifier/limiter.

The differential current-mode outputs of all nine detector cells are added together and fed to a 'current mirror' output stage, which effectively converts them into a direct current.

Because of the combination of cascaded gain and limiting in the amplifiers (plus an internal offset compensation loop), the amplitude of this output current is proportional to the logarithm of the AC input voltage. This holds true over an input range of just on 100dB, from about –93dBV (22.4µV) up to +7.0dBV (2.24V). This logarithmic relationship is linear to within ±0.3dB over most of the range.

The output current (I_{OUT}) increases at a slope of very close to 2µA per dB increase in AC input level, and when this current passes through a 12.5kΩ load resistor inside the chip, the result is a DC output voltage of 25mV/dB. This slope can be fine-tuned using an external adjustable resistor in parallel with the 12.5kΩ internal resistor.

The 'set intercept' (SI) pin allows you to adjust the DC offset in the output current mirror, which sets the effective zero-level point of the chip's output current and voltage; ie, the origin from which the output slope rises.

You can now fit the remaining SMD passives, plus the two SMD diodes, ensuring their cathode stripes face as shown in Fig.5.

Next, fit transistor Q1. It has three pins, so its orientation should be obvious. Make sure its leads are sitting flat on the PCB before you solder it in place.

The last SMD component is L2, which is quite large. Spread a thin smear of flux paste on both pads before you start.

You will need a hot iron to form good solder joints due to the thermal masses of both the PCB and the part. Make sure you add enough solder and heat it long enough to form good fillets.

Through-hole parts

Before proceeding, we need to wind choke L1 and transformer/common-mode choke T1. These are both wound on 5mm-long ferrite beads, using 0.25mm-diameter enamel-coated copper wire.

L1 has three single turns, while T1 has three bifilar turns, wound by first folding a 200mm length of the wire in two, and then using the 'doubled pair' to wind their three turns together.

Once both chokes are wound, cut off the wire ends about 8mm from the ends of the ferrite beads, scrape off about 4mm of the enamel and then lightly tin the wire ends so they will be easy to solder into the PCB pad holes.

Just before you solder in the four wires for T1, use your DMM to make sure that the wire pairs do not 'cross over'; the left-most upper and lower wires should be joined together, as should the right-most upper and lower wires.

You can now proceed to fit the remaining through-hole parts. Start with diode D1 (as usual, be careful with its orientation). It's then a good idea to install the six PC pins, if you are going to use them. These make it easier to use clip leads to connect your DMM to the board during testing and calibration. These are for TPGND, TP2.5V, TP5V and TP1-TP3.

Next, mount the reed relay, again taking care with its polarity. Follow with the two multi-turn trimpots, which are different values (so don't get them mixed up), followed by the 4-pin header for CON3, the 3-pin header for JP1 and the 2-pin header used to facilitate the connection of LED1.

Now is also a good time to install the 1µF through-hole capacitor, near IC4.

Before you mount the Nano board, you will need to fit a short length of wire shorting out its onboard diode D1, on the underside; see the sidebar photo and text for an explanation of why this is necessary and how to do it.

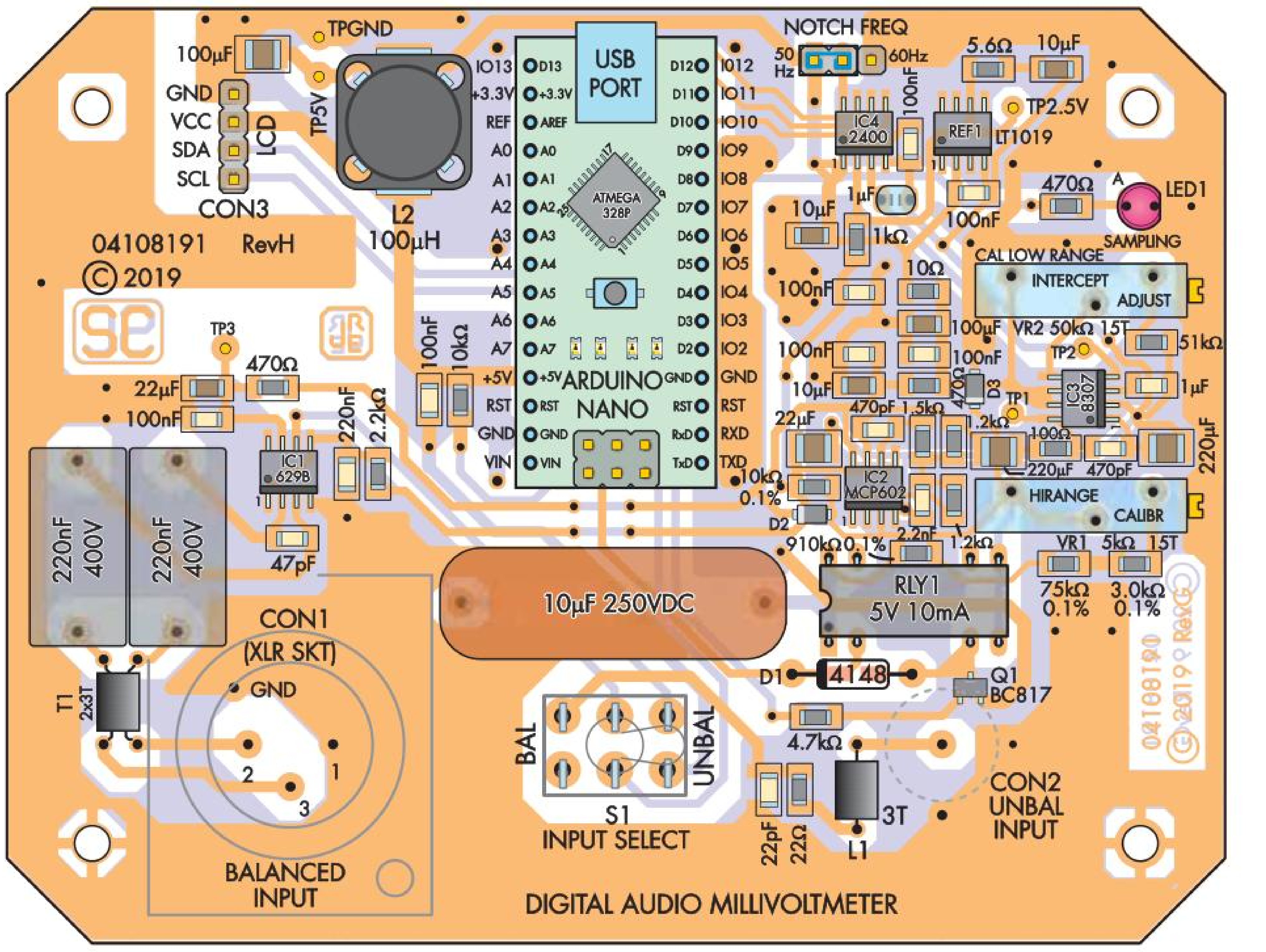
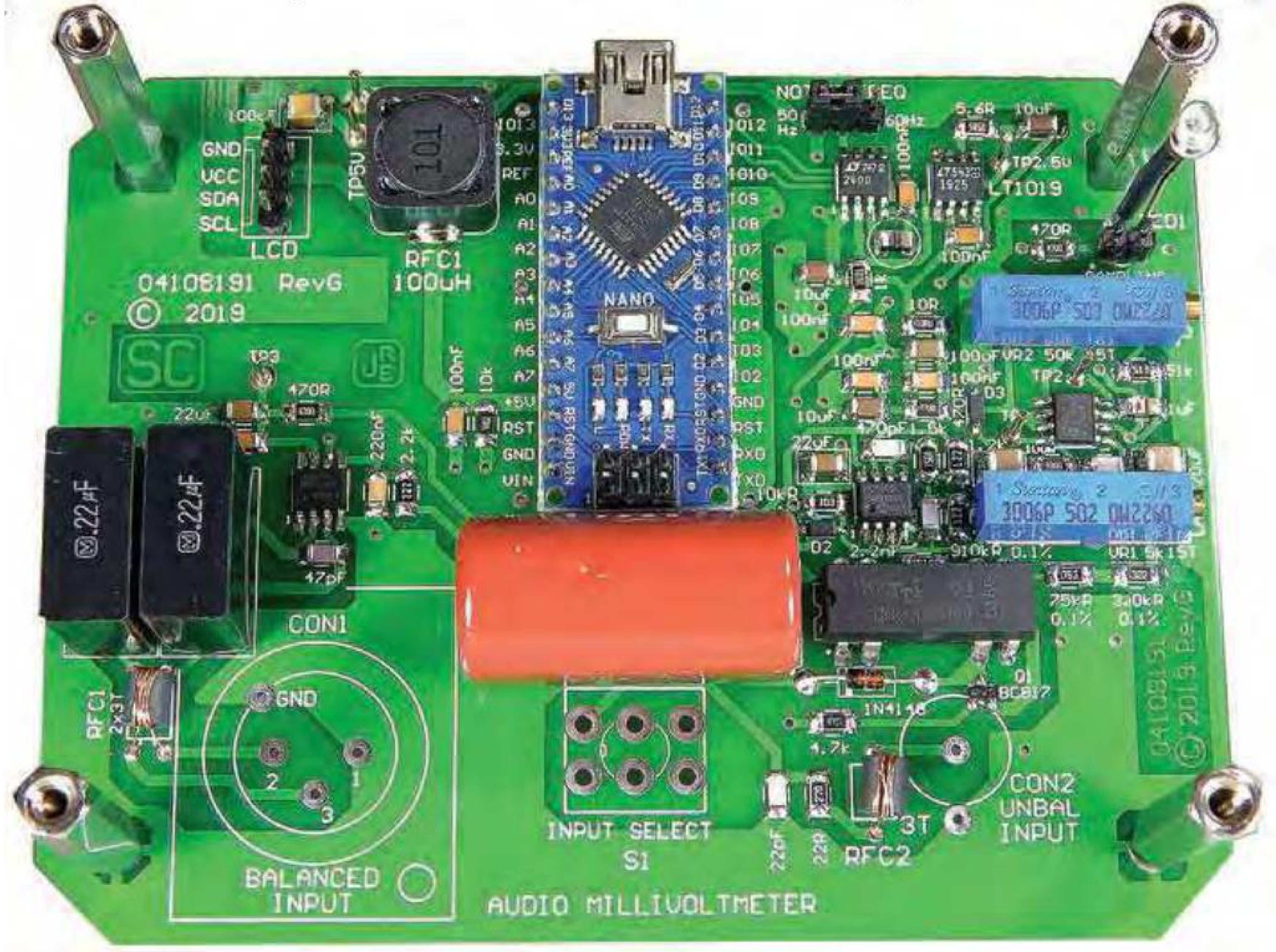

TPGND
100µF
GND
VCC
SDA
SCL
LCD
CON3
TP5V
L2
100µH
04108191 RevH
© 2019
USB PORT
ATMEGA 328P
ARDUINO NANO
NOTCH FREQ
50 Hz
60Hz
5.6Ω
10µF
IC4 2400
REF1 LT1019
TP2.5V
470Ω
LED1
A
1µF
10µF
1kΩ
100nF
10Ω
CAL LOW RANGE
SAMPLING
INTERCEPT
ADJUST
VR2 50kΩ 15T
51kΩ
TP2
IC3 8307
TP1
D3
22µF
470pF
1.5kΩ
1.2kΩ
100Ω
220µF
HIRANGE
CALIBR
IC2 MCP602
D2
910kΩ 0.1%
2.2nF
VR1 5kΩ 15T
75kΩ 0.1%
3.0kΩ 0.1%
RLY1 5V 10mA
10µF 250VDC
TP3
470Ω
22µF
100nF
IC1 6298
220nF
2.2kΩ
10kΩ
220nF 400V
47pF
CON1 (XLR SKT)
GND
1
2
3
T1 2x3T
BALANCED INPUT
D1
4148
Q1 BC817
4.7kΩ
BAL
UNBAL
S1
INPUT SELECT
22pF
22Ω
L1
3T
CON2 UNBAL INPUT
DIGITAL AUDIO MILLIVOLTMETER

Fig.5: this PCB overlay diagram (and photo below) shows where the components are mounted on the PCB, including the prebuilt Arduino Nano microcontroller module. Most of the components are larger SMD types which are not difficult to hand-solder. Some components, such as CON1, CON2 and S1 are mounted on the lid (front panel) and wired back to the board using short leads.



Now solder the Arduino Nano module to the rows of pads on the board, with its USB connector over the outside edge. Make sure it's pushed all the way down before soldering; it's a good idea to solder two diagonal pins first, check that it's flat and then solder the rest.

Finish up by mounting the three large capacitors.

The final step at this stage is to solder the leads of LED1 to the pins of the 2-pin header fitted to the PCB, taking care to connect them to the correct pin (the longer anode pin goes to the inner pin marked 'A'). The leads should be soldered to the pins so that the underside of the LED's body is 28mm above the top of the PCB.

Your *Meter's* PCB assembly should now be complete and ready to be fitted into the box, once it has been prepared.

Before you do so, though, plug the 4-pin female socket onto CON3 and place the shorting block in the correct position on JP1, to suit your local mains frequency.

Preparing the box

Most of the holes you'll need to drill or cut in the box are in the lid, which becomes the *Meter's* front panel.

There are only three holes to be cut in the base of the box: two circular holes in the right-hand end for access to trimpots VR1 and VR2, and one rectangular hole in the centre of the box rear to allow access for the power/PC USB connector.

You'll find the location and sizes of all of these holes in the two drilling diagrams (Figs.6 and 7). Most of the holes are circular and can be drilled, although the 23mm-diameter hole for XLR connector CON1 is best made using either a hole saw or by drilling a circle of small holes and then cutting between them using either a rat-tailed file or jeweller's saw.

Parts list – Digital Audio Millivoltmeter

1 119 × 94 × 57mm diecast aluminium box [Jaycar Cat HB-5064 or similar]
1 double-sided PCB, 109 × 84mm, code 04108191 (RevH)
1 Arduino or Duinotech Nano MCU module
1 USB Type-A to mini Type-B cable
1 16x2 backlit alphanumeric LCD module with I^2C serial interface
  [eg, Silicon Chip Online Shop Cat SC4198 or similar]
1 panel-mount miniature DPDT toggle switch (S1) [Jaycar ST035, Altronics S1345]
1 panel-mount 3-pin female XLR connector (CON1) [Jaycar PS1930, Altronics P0804]
1 panel-mount BNC socket (CON2)
1 4-pin header, 2.54mm pitch (CON3)
1 4-pin female header socket, 2.54mm pitch (to connect LCD module)
1 2-pin header, 2.54mm pitch (for LED1)
1 3-pin header with jumper shunt (JP1)
1 SPST DIL reed relay with 5V/10mA coil (RLY1) [Jaycar Cat SY-4030 or similar]
2 5mm-long ferrite beads, 4mm outer diameter (L1,T1)
  [Jaycar Cat LF-1250 or similar]
1 300mm length of 0.25mm-diameter enamelled copper wire (for L1 and T1)
1 100µH SMD RF inductor (L2) [Jaycar Cat LF-1402 or similar]
4 25mm-long M3 tapped spacers
4 6mm-long untapped spacers
8 12mm or 15mm-long M3 panhead machine screws
2 9mm-long M3 countersunk head machine screws
2 M3 hex nuts and star lockwashers
4 16mm or 20mm-long M2.5 countersunk head machine screws
4 9mm-long untapped spacers, >2.5mm inner diameter
4 M2.5 hex nuts
6 PCB pins (optional; for TPGND, TP2.5V, TP5V and TP1-TP3)

Semiconductors
1 AD629BRZ high common-mode-voltage difference amplifier, SOIC-8 (IC1)
1 MCP602-I/SN dual rail-to-rail input/output op amp, SOIC-8 (IC2)
1 AD8307ARZ logarithmic amplifier/detector, SOIC-8 (IC3)
1 LTC2400CS8#PBF 24-bit ADC, SOIC-8 (IC4)
1 LT1019ACS8-2.5#PBF precision 2.500V voltage reference, SOIC-8 (REF1)
1 BC817-40 NPN transistor, SOT-23 (Q1)
1 3mm red LED (LED1)
1 1N4148 silicon small-signal diode (D1)
2 1N5711W-7-F schottky diodes, SOD-123 (D2,D3)

Capacitors (all SMD ceramic, 3216/1206 size unless otherwise stated)
2 220µF 6.3V X5R, SMD 3226/1210 size
2 100µF 6.3V X5R
2 22µF 10V X5R
3 10µF 16V X7R
1 10µF 250VDC metallised polypropylene,radial leaded [Panasonic ECQ-E2106KF]
1 1µF 50V through-hole ceramic or MKT
1 1µF 16V X7R
2 220nF 275VAC metallised polypropylene, radial leaded [Panasonic ECQ-U2A224ML]
1 220nF 16V X7R *(Code 220, 0.22 or 220n)*
7 100nF 16V X7R *(Code 100, 0.1 or 100n)*
1 2.2nF 16V X7R *(Code 2.2, .022 or 2n2)*
2 470pF 100V C0G/NP0 *(Code 470, .0047 or 470p)*
1 47pF 100V C0G/NP0 *(Code 47, .00047 or 47p)*
1 22pF 250V C0G/NP0 *(Code 22, .00022 or 22p)*

Resistors (all SMD 1% 0.25W, 3216/1206 size unless otherwise stated)
1 910kΩ 0.1% 1 75kΩ 0.1% 1 51kΩ 1 10kΩ 1 10kΩ 0.1%
1 3.0kΩ 0.1% 1 4.7kΩ 1 2.2kΩ 1 1.5kΩ 2 1.2kΩ
1 1kΩ 1 100Ω 3 470Ω 1 22Ω 1 10Ω 1 5.6Ω
1 5kΩ multi-turn horizontal trimpot (VR1)
1 50kΩ multi-turn horizontal trimpot (VR2)

The best plan for cutting the 65 × 15mm rectangular hole for the LCD screen is to drill a 6mm diameter hole inside each corner, to allow you to use a small metal-cutting jigsaw to cut along each side.

Then you can tidy up the edges using a small file.

For the rectangular hole in the rear of the box, I first drilled a 9mm diameter hole in the centre, then used jeweller's files to expand it out into the final rectangular shape.

Once all of the holes have been made, remove all burrs from the inside and outside of each hole using one or more small files.

As a final step in preparing the box for assembly, you should fit a professional-looking panel on the lid.

We have produced a front panel artwork for this project, which can be downloaded from the October 2020 page of the *PE* website as a PDF file. You can then print, laminate and attach it to the lid using thin double-sided adhesive tape or a smear of silicone sealant.

The final step is to cut out the holes in the dress front panel to match those in the lid itself, using a sharp knife.

Final assembly

Glue an 80 × 40mm rectangle of 0.5mm-thick clear plastic sheet to the rear of the lid, just behind the LCD window. This is to keep dust out and protect the LCD screen from accidental scratches. It can be cut from a clean takeaway container lid or similar.

Then mount the LCD screen to the underside of the lid using four 16mm-long M2.5 countersunk-head screws with four 9mm-long untapped spacers and four M2.5 nuts, as shown in Fig.8. Next, fit XLR connector CON1 to the lid using two 9mm-long countersunk-head M3 screws with lock washers and nuts on the rear.

After this, fit BNC connector CON2 using its matching lock washer, solder lug and nut, then input selector switch S1.

To ensure that the switch is fixed in place horizontally, you can drill a small blind hole in the rear of the lid to accept the spigot on the edge of the switch's flat washer.

Now up-end the lid/front panel and solder stiff wire leads to the rear lugs of CON1, CON2 and S1. These don't have to be very long; just long enough to pass down through their matching holes in the PCB when it's fitted.

The only one that needs special treatment is that for CON2, which should ideally be made using a 25mm length of shielded microphone cable.

Take care when separating the screen wires at each end, to prevent accidental shorts.

Once these extension leads have been fitted, you are ready to mount the PCB to the rear of the lid/front panel.

The PCB is mounted using four 25mm-long M3 tapped spacers, together with four 6mm-long untapped spacers, as shown in Fig.8. First attach all four pairs of spacers to the corners of the PCB, using 12mm-long M3 screws passing up through the PCB and the untapped spacers, and then into the 25mm tapped spacers.

The complete PCB-and-spacers assembly is then attached to the rear of the lid/front panel, using four 12mm-long M3 screws.

While doing this, ensure that the extension wires from CON1, S1 and CON2 pass through their matching holes in the PCB. And before you finally tighten up the screws, make sure that the body of LED1 is protruding through its matching hole in the front panel.

Now solder the ends of the extension wires from CON1, S1 and CON2 to their matching pads on the rear of the PCB.

If all has gone well so far, you should find that the pin ends of the 4-pin SIL header fitted to the end of the LCD module are now very close to those of the socket plugged into CON3.

You should only need to bend the module's header pins down slightly to meet the pins from CON3's socket, and then you can solder them together.

Your *Meter* is now complete, apart from the final fitting of the front panel assembly into the box.

But before you do this, it's a good idea to load the *Meter's* firmware sketch (program) into the Arduino Nano. This is done using the Arduino IDE, running on a suitable PC, with the *Meter* connected to a USB port of the PC via a standard USB Type-A to mini Type-B cable.

Programming the Meter

The firmware program to be loaded into the Meter's Arduino Nano is called **AudiomVmeterMk2_sketch.ino**, which you can download from the October 2020 page of the *PE* website. Save it in a folder where you'll be able to find it later.

Now is also a good time to make sure that you have the latest Arduino IDE (integrated development environment) installed. If not, you can get it from: **www.arduino.cc/en/main/software**

This software allows you to compile and upload the code to the Arduino board.

Plug the *Meter* into your PC, and its LCD backlight should light up, showing that the *Meter* is receiving 5V power.

Assuming that you are running Windows, open the Control Panel and select 'System and Security' and then

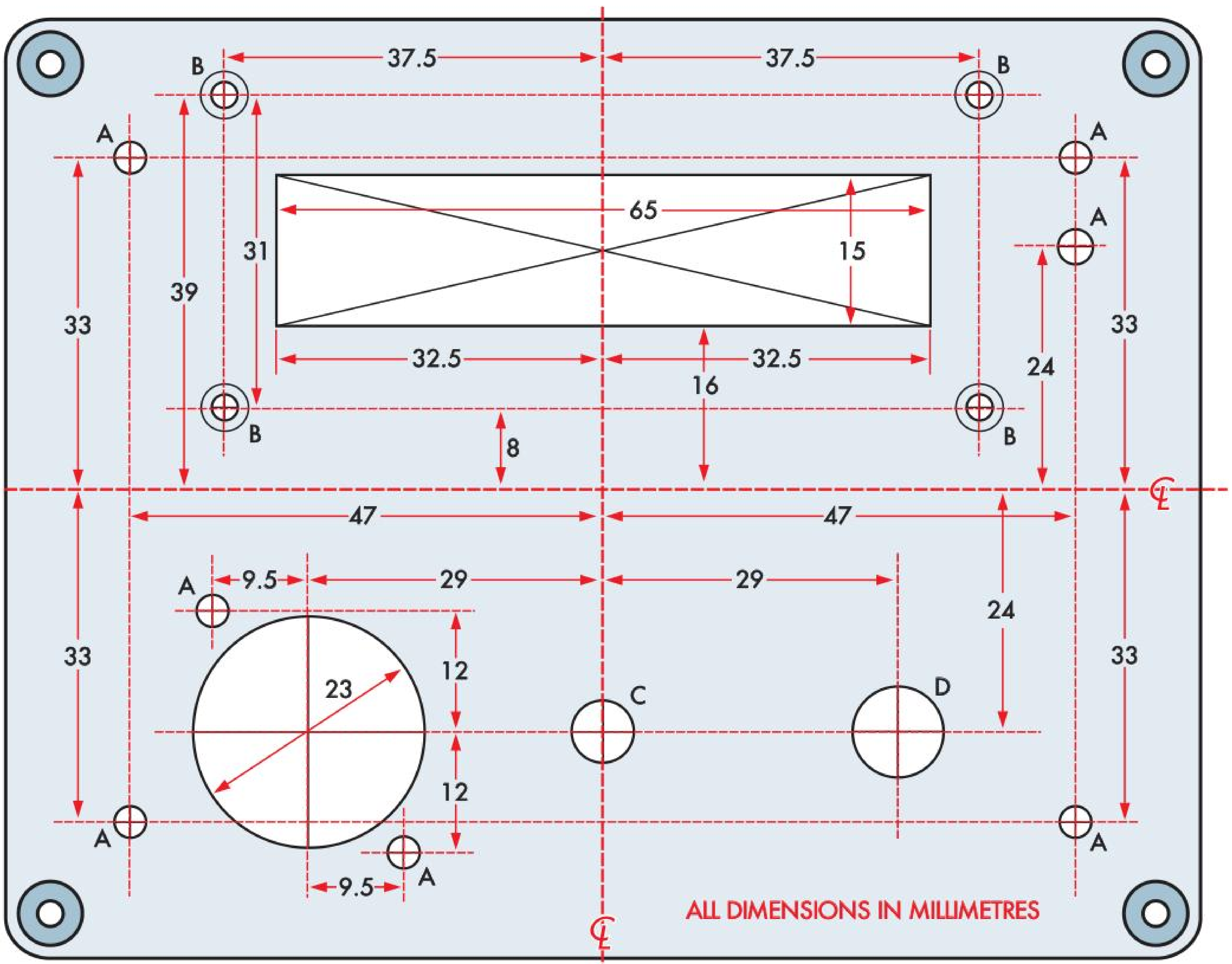


Fig.6: most of the holes that need to be made in the case go in the lid. Holes A are 3mm diameter, B are 2.5mm, C 6.5mm and D 9mm. You'll probably need a hole saw to cut the 23mm, although you could use a 20mm stepped drill bit and then enlarge to 23mm with a large tapered reamer. Note that holes 'B' need to be countersunk after being drilled. See the text for suggestions on how to make the large rectangular cut-out.

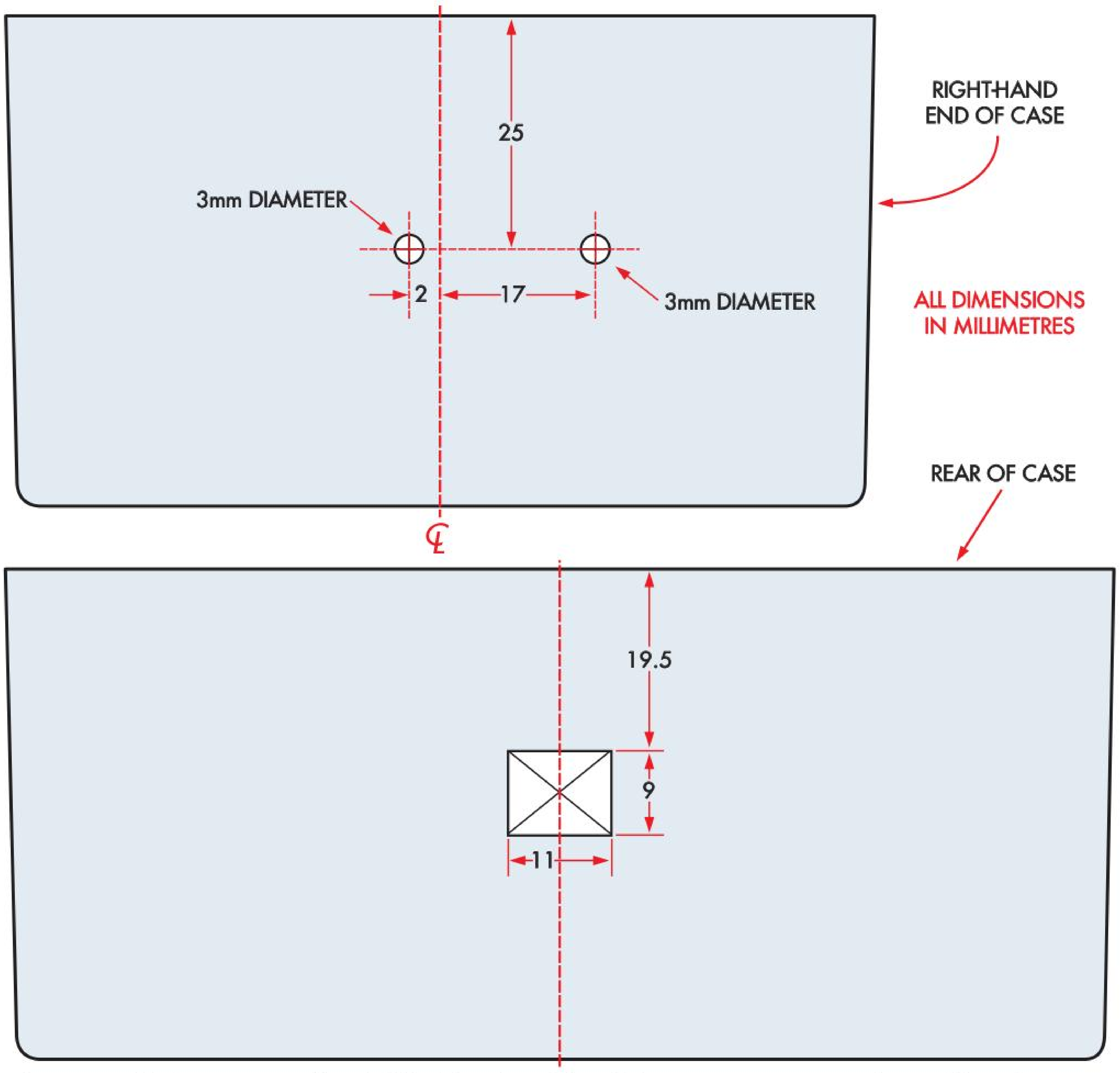
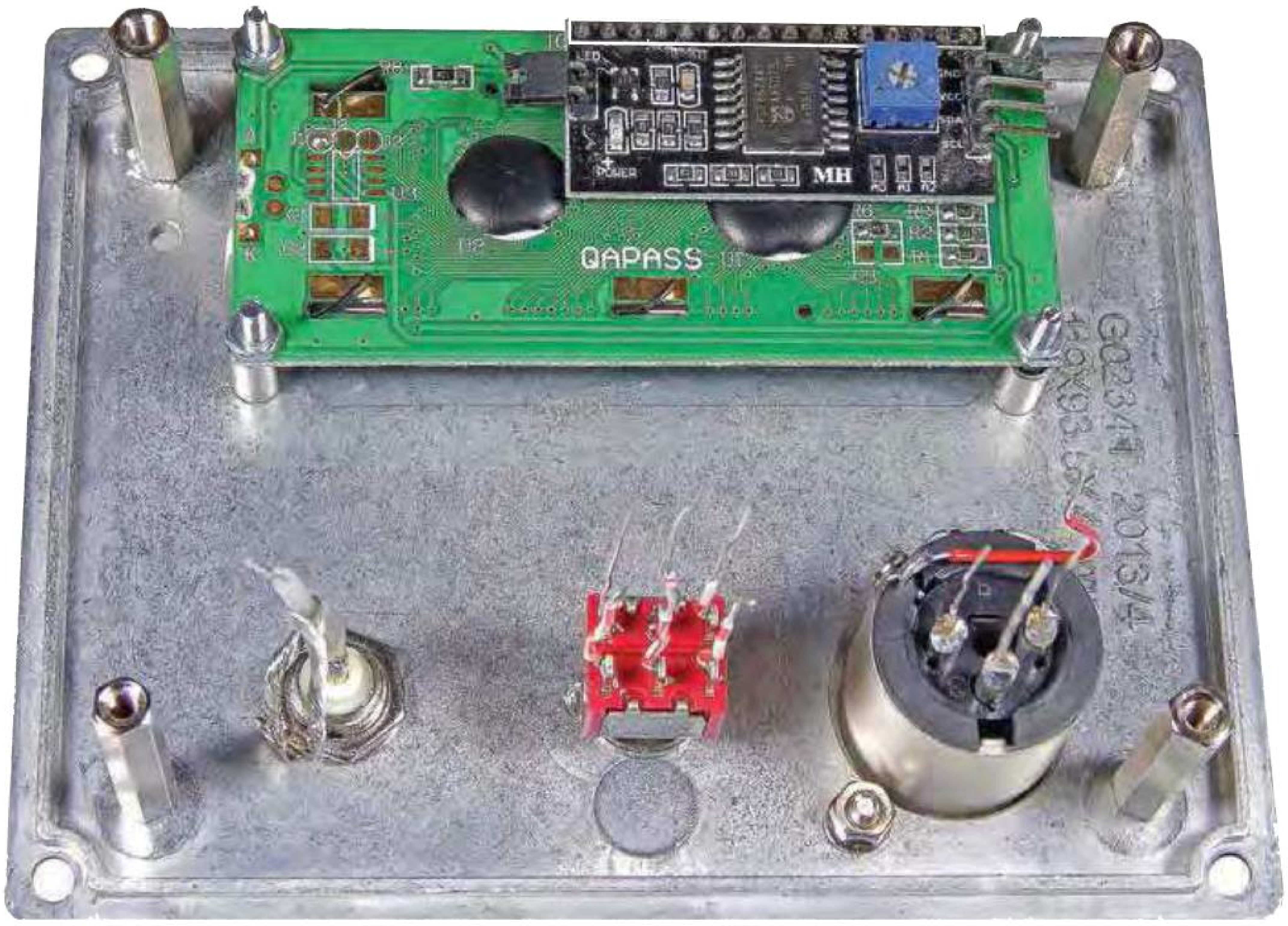


Fig.7: two holes need to be drilled in the side of the case to access the calibration potentiometer screws, while a small rectangular cut-out on one of the long sides provides access to the USB socket, both for power and optionally for logging measurements to a PC.



The pre-assembled display PCB mounts so that the LCD lines up with the cutout in the lid (which becomes the front panel). Here we also show the four mounting pillars and the input select switch along with the XLR and BNC sockets, with their connecting wires already soldered in place and ready to connect to the main PCB.

'Device Manager'. This should allow you to see the Virtual COM Port that the *Meter* has been allocated. It should also allow you to set the baud rate for communication with the *Meter*. Set it to 115,200 bps.

Now start up the Arduino IDE and load the sketch that you downloaded earlier. In the IDE's Tools menu, set the Board selection to 'Arduino Nano' and the Processor to 'ATMega328P (Old Bootloader)', then set the COM Port to whichever one your *Meter* is connected to, as determined earlier.

Open the sketch and in the Sketch menu, click on 'Verify/Compile'.

When you get the 'Compiling Done' message, go to the Sketch menu again and click on 'Upload'. The compiled sketch should then be uploaded into the Nano MCU's Flash memory.

After a few seconds, the *Meter* should start up, giving you a brief message on the LCD announcing itself. It will then start sampling from whichever input S1 is set to select.

At this stage, the *Meter* may not be giving sensible readings, since it has yet to be calibrated. But you can check the various DC voltages on the PCB test points.

For example, you should find a voltage very close to 5V between TP5V and TPGND, while the voltage at TP2.5V should read 2.500V with respect to TPGND.

If those check out, you can now install your *Meter* in its box, by lowering it in and then screwing the lid with the four M4 countersunk screws supplied with it.

Calibration

For accurate results, your *Meter* must be calibrated. You'll need access to an audio oscillator or a function generator, together with a DMM capable of making accurate and reasonably high-resolution AC voltage measurements in the range from 500mV to 10V (RMS).

Power up the audio oscillator or function generator and set it to provide a 1kHz signal with an amplitude of 600mV RMS (1.697V peak-to-peak).

Check this level using your DMM, and adjust the generator if necessary. Then power up the *Millivoltmeter* and connect the oscillator's output signal to the *Meter's* unbalanced input (CON2), with S1 set appropriately.

After a few seconds, the *Meter* should show a stable reading in both millivolts and dBV, with the legend '(L)' at lower right. This indicates the *Meter* has switched to its lower range.

At this stage, the reading will probably differ a little from the correct value of 600mV and –4.437dBV. So use a small screwdriver or alignment tool to adjust trimpot VR2 (INTERCEPT ADJUST), to bring the reading as close as possible to that correct value. This calibrates the *Meter's* low range.

The next step is to calibrate the *Meter's* high range. Change the output level of the audio oscillator or function generator to 10.000V RMS (28.28V peak-to-peak), checking this using your DMM again.

If your oscillator or function generator can't provide an output that high (which is quite common), you may have to use a small amplifier to boost its output.

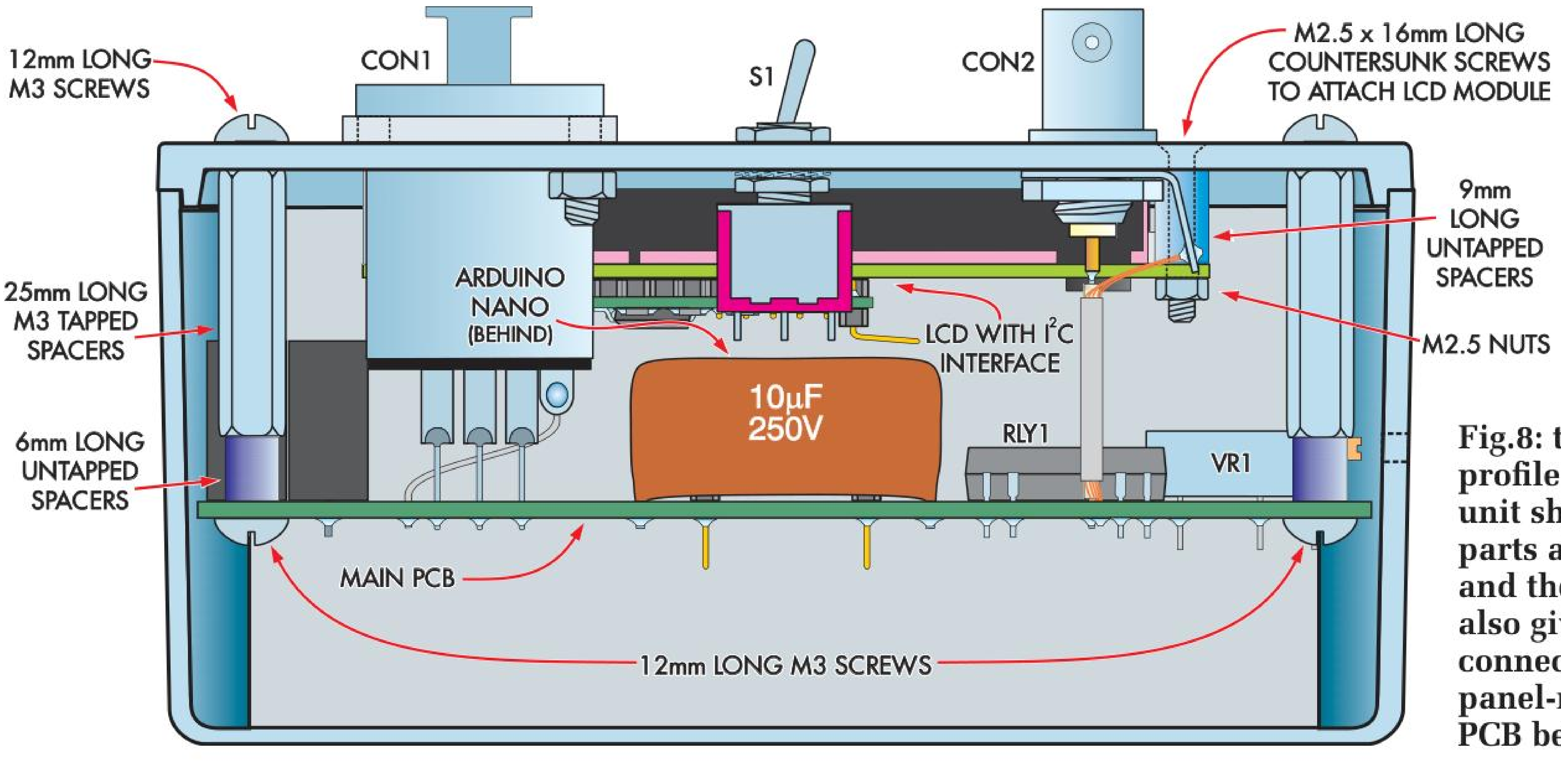
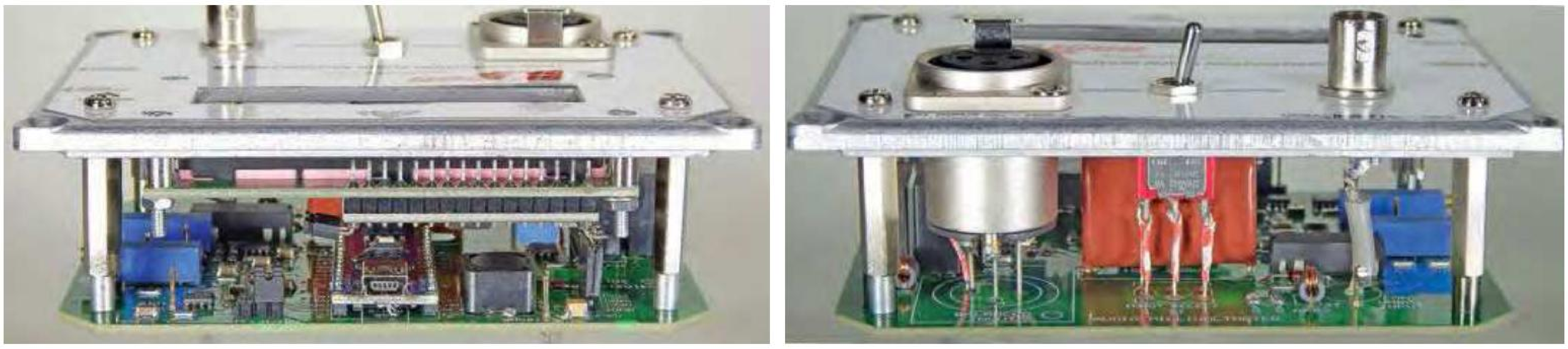

12mm LONG M3 SCREWS
CON1
S1
CON2
M2.5 x 16mm LONG COUNTERSUNK SCREWS TO ATTACH LCD MODULE
9mm LONG UNTAPPED SPACERS
25mm LONG M3 TAPPED SPACERS
ARDUINO NANO (BEHIND)
LCD WITH I²C INTERFACE
M2.5 NUTS
10µF 250V
RLY1
VR1
6mm LONG UNTAPPED SPACERS
MAIN PCB
12mm LONG M3 SCREWS

Fig.8: this 'cut-away' side profile view of the assembled unit shows how the various parts attach to each other and the back of the lid, and also gives you an idea of the connections needed from the panel-mounted parts to the PCB below.



'Left and right' views of the assembled project immediately before it is mounted in the diecast case. The input sockets and selector switch are all connected to the PCB via short lengths of either tinned copper wire or, in the case of the BNC socket (CON2), shielded cable. The photo at right compares with the diagram on the opposite page.

An amplifier capable of doing just that, very accurately, is described starting on page 29 of this issue.

Now connect the oscillator's output signal to the *Meter's* unbalanced input (CON2) again, and after a couple of seconds, the *Meter* should display a new reading.

This time, the legend at the end of the lower line should read '(H)', to show that it has now switched to the higher range.

The new reading is likely to be fairly near the correct value of 10.000V and 20.00dBV, but not spot-on. Correct it by adjusting trimpot VR1 (CALIBRATE HI RANGE).

Once this has been done, your new *Digital Millivolt/Voltmeter* is calibrated and ready for use.

Logging measurements

All you need to do to log measurements to your PC is open up the Arduino Serial monitor, using the same settings as described above for programming the Nano.

With the unit connected to your PC, each time it takes a measurement, it will also be written to the serial monitor.

When you have finished, you can save the log for later analysis (for example, using mathematical functions in a spreadsheet).

This view of the right end of the PCB shows the two 15-turn trimpots, VR1 (left – 5kΩ) and VR2 (right – 50kΩ) which are used to set the HIGH range calibration and intercept adjust, respectively (see text). These pots line up with access holes drilled in the end of the case.

Ensuring that a low-cost Arduino Nano works reliably

There are Arduino Nanos... and there are Arduino Nanos!

During the development of this project, we discovered on two occasions that the 'El cheapo' Arduino Nanos had started to malfunction.

In both cases, diode D1 in the Nano's power supply had 'blown' and changed into a high resistance, lowering the supply voltage to less than 2.8V.

This diode (an SS1 or an MBR0520) is not really required when the Nano is powered from USB. It's purely to protect the USB port of the PC when the Nano is powered via a higher voltage supply fed directly into its Vin pin.

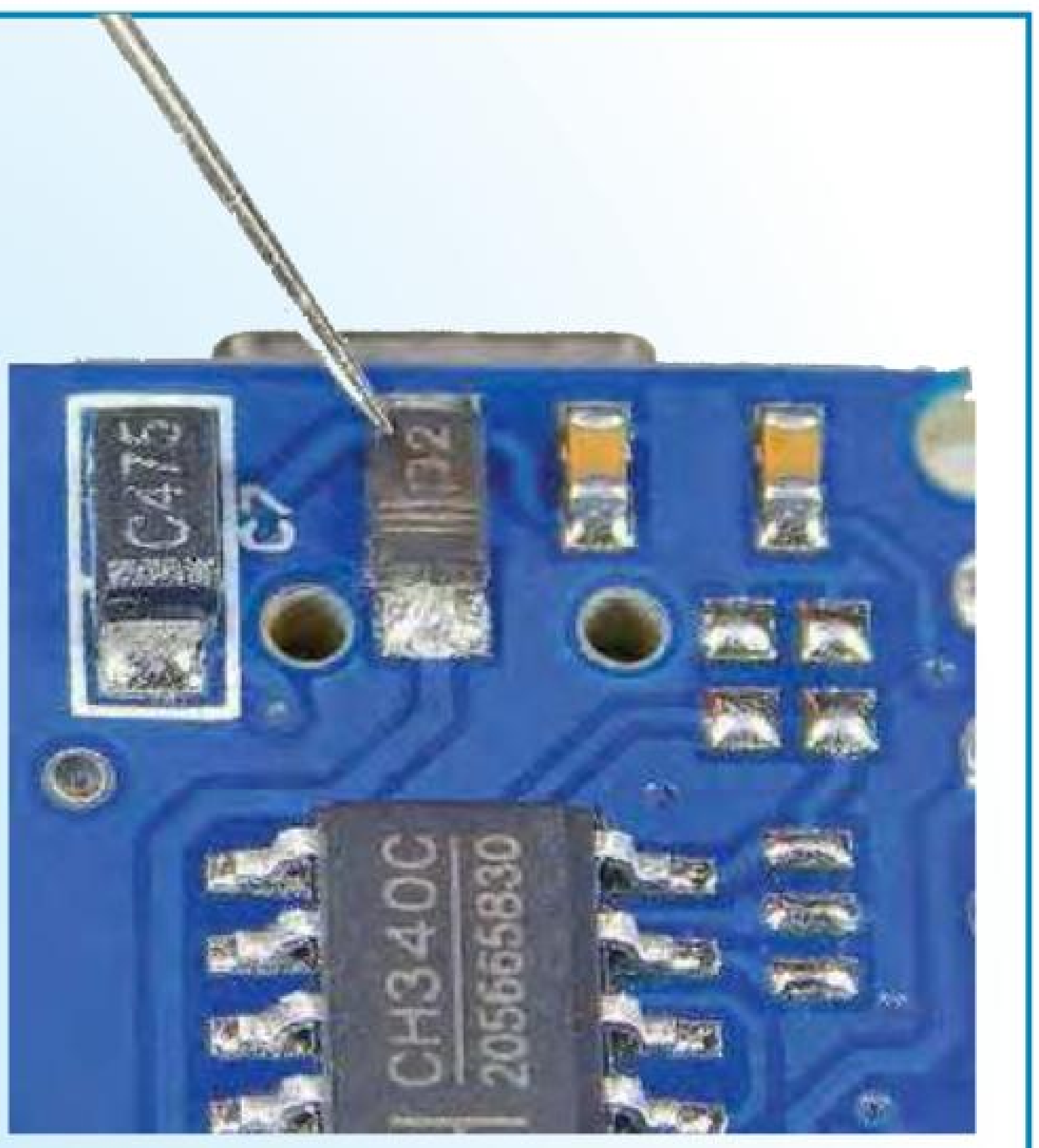
Since the Nano and its associated circuitry (here, the Millivoltmeter) are always going to be powered from the USB connector, there's no reason why the diode can't be simply shorted out, to ensure reliable operation.

The problem is that the diode is fitted to the underside of the Nano's tiny PCB. This makes it quite inaccessible if the Nano has already been fitted to your Meter's main PCB. In fact, I had to virtually destroy the first Nano to remove it from the main PCB to get at the blown diode.

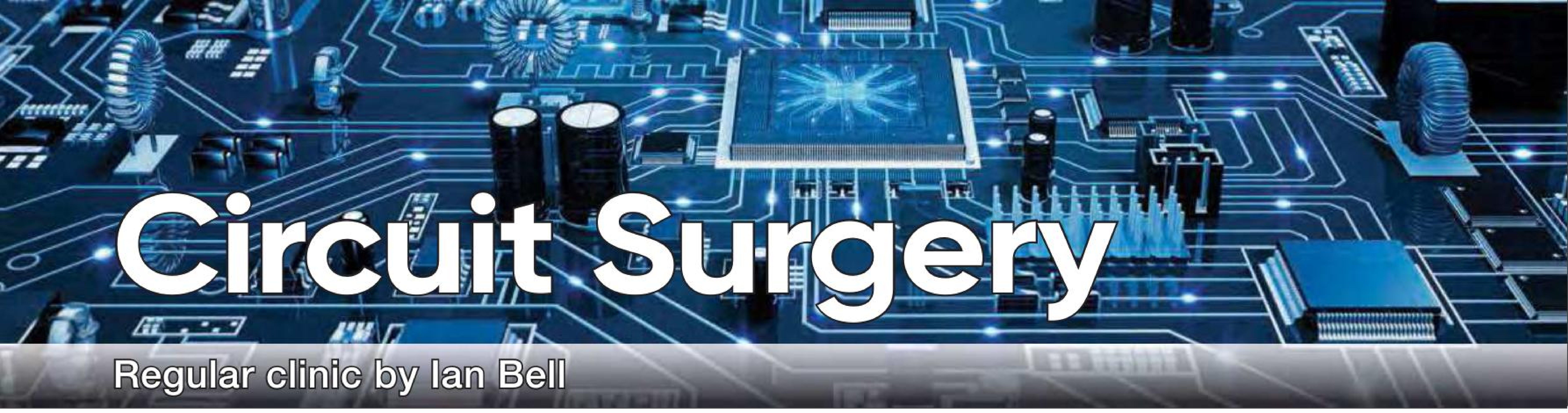
So we suggest that if you are going to be using a low-cost Nano in your Millivoltmeter, you should first short out D1 with a short length of wire, before mounting it on the main PCB.

This should ensure reliable operation and avoid the need for surgery at a later stage.

The photo at right shows where D1 is located, just below the Mini USB connector. The diode is usually marked 'B2', although on the one in the photo it looks more like 'D2' because there's a tiny crater in the middle of the B where the smoke came out.



It's quite easy to short out the diode with a short length of tinned copper wire, bent into a tiny inverted 'U'. If you use the same soldering iron you use to fit SMD components, it can be done quite quickly if you're careful. Just make sure that the wire link doesn't protrude upwards very far, or it might touch the top copper of your main PCB when the Nano is mounted on it.



Circuit Surgery

Regular clinic by Ian Bell

Rail-to-rail and single-supply op amps

This month's topic is inspired by a question posted on the EEWeb Forum by Alexandru Radu, who wrote: 'I want to design a rail-to-rail input and output op amp and I want to be sure I understand what it really does, so I will give a few examples. First of all, from what I understood, a regular op amp can't really reach the maximum swing, but a rail-to-rail op amp can, even surpass it, but for simplicity let's say it is exactly the supply voltage.

Let $V_{DD} = 5V$ and $V_{SS} = -5V$, and set up my op amp as an inverting amplifier. If my input signal is 1V, R2 = 5kΩ and R1 = 1kΩ, then my output will be exactly 5V, right? For a usual op amp it could be something like... 4.9 or 4.95... Is this correct? Thank you!'

Alexandru's overview is basically along the right lines but there are plenty of details to discuss – we will look at rail-to-rail op amps and at the related topic of single-supply op amp circuits.

Rails and swings

Before getting into the details it is worth defining our terms and context. The word 'rail' refers to the power supply connection to a circuit, or the voltage of that power supply. Modern electronic systems often have many different supply voltages for different subsystems, but when we focus on a particular circuit within a system, such as an amplifier, we typically have two supply rails. A key aspect of some circuit designs is the relationship between the signal-voltage range and the supply voltages. The signal voltage range handled by a circuit is often referred to as the voltage 'swing', and is commonly described with respect to the supplies; for example, 'the output can swing to within 1V of the rails'.

Many circuits can only work with input signal voltages which fall inside the range defined by the two supply voltages and are also only able to output voltages within that range. If inputs exceed the supply range (in either direction) the circuit may not operate correctly or may be damaged, although it is not uncommon for op amps to work with inputs a few hundred millivolts outside the supply range. For outputs, it is often physically impossible for the circuit to output voltages outside the supply range. There are of course exceptions, for example DC-DC converters output higher voltages, but our discussion here is focused on analogue signal-processing circuits built with op amps (such as amplifiers and filters) where the within-supply-range restrictions usually apply to output voltages.

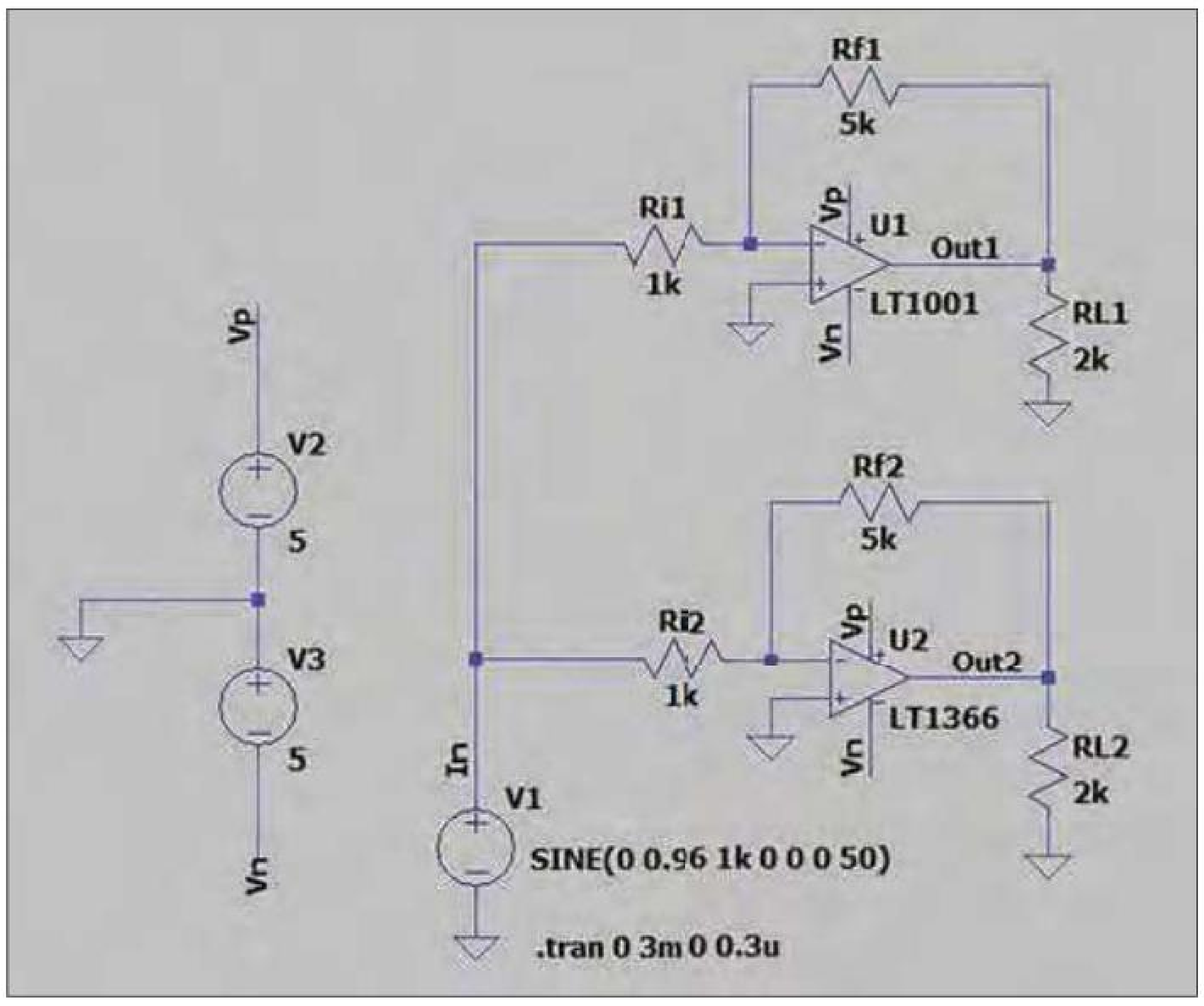


Fig.1. Simulation schematic to obtain example waveforms. The LT1001 is a 'Precision Op Amp' and the LT1366 is a 'Precision Rail-to-Rail Input and Output Op Amp'.

As Alexandru indicates, the term 'rail-to-rail' can apply to both the input and output voltages of op amps. These are separate capabilities – rail-to-rail input does not imply rail-to-rail output, but many op amps branded as rail-to-rail cover both input and output. These voltages apply to the op amp itself – the input/output voltages of the circuit as a whole may be different. For op amp inputs it is the common-mode voltage which is of importance – more on this later.

In many situations the limited signal range of standard op amps is not an issue because the signals never go to the levels which cause problems. However, if operation close to the supplies is required then a special effort must be made to design circuits which can operate over a wider signal – so we have rail-to-rail op amps. A question that may occur at this point is, why aren't all op amps designed to be rail-to-rail from the start? One answer is that it is simply more difficult, and, in the past, it was less likely to be an issue. Over the last two or three decades, supply voltages have tended to reduce due to the effects of advances in semiconductor technology. If an op amp is limited to signals to within 1V of the supply this is much more likely to be a problem with a 3.3V supply than a 15V one. Furthermore, as is often the case in engineering, improving one aspect of a device may degrade some other characteristics. If rail-to-rail is not a requirement then a suitable standard range op amp may provide better performance.

Rail-to-rail outputs

Alexandru mentions some example output voltages – although 4.9V on a 5V supply is more rail-to-rail than a 'usual' op amp. In general, rail-to-rail output does not mean fully to the supply voltages – it is more of a marketing term to indicate capabilities beyond the standard, not an exact specification. Typically, for BJT (bipolar junction transistor) op amps, rail-to-rail outputs can go to within a collector-emitter saturation voltage (V_{CEsat}) of the supply. V_{CEsat} is dependent on the transistor's collector current and hence the op amp's output current. For moderate currents (in the mA range) V_{CEsat} is typically 100 to 300mV, so that is around 4.7 to 4.9V maximum for 5V a supply. For non-rail-to-rail op amps the output limits are typically 1V to 2V away from the supply. For example, the outputs swing for the venerable LM741 is ±12 to ±14V on a

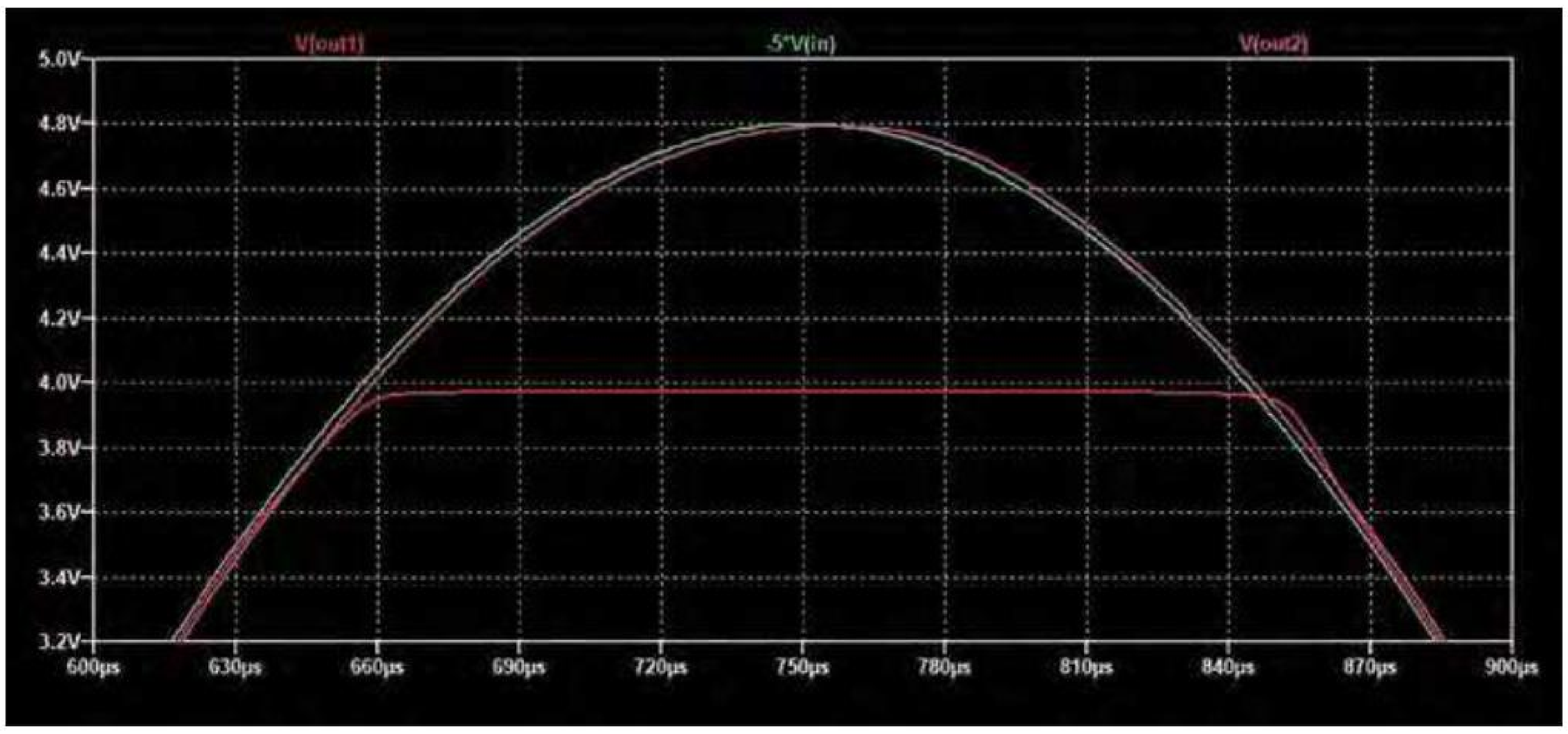


Fig.2. Positive peak of output waveforms from circuit in Fig.1 with Vin=0.96V. The green trace is ideal (–5 times the input voltage); the red trace is a standard op amp; the magenta trace is a rail-to-rail op amp swinging to 200mV below the 5V supply without distorting the signal.

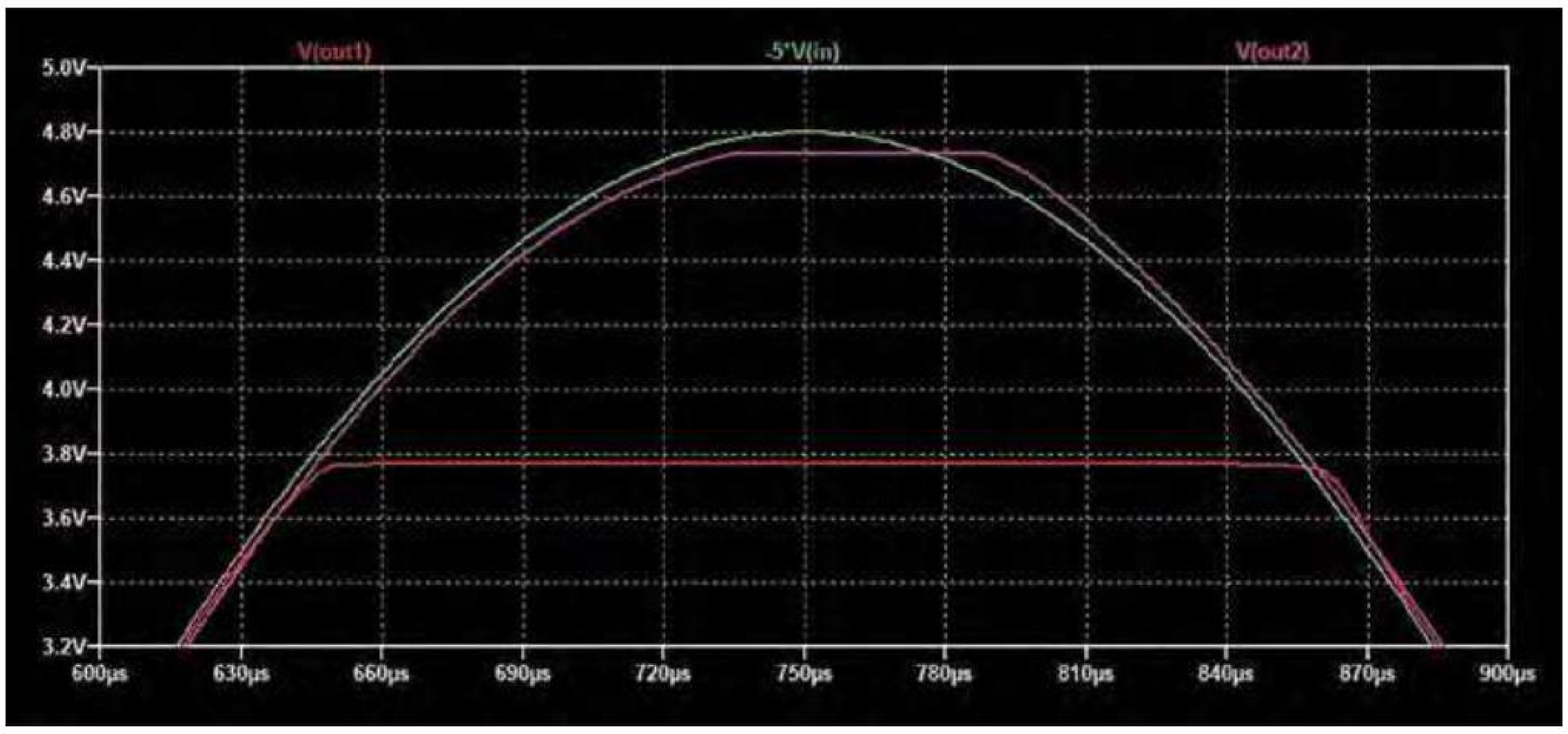


Fig.3. Same simulation as in Fig.2 but with the load resistors reduced to 500Ω. Both outputs peak at a lower voltage and there is now significant distortion from the rail-to-rail op amp.

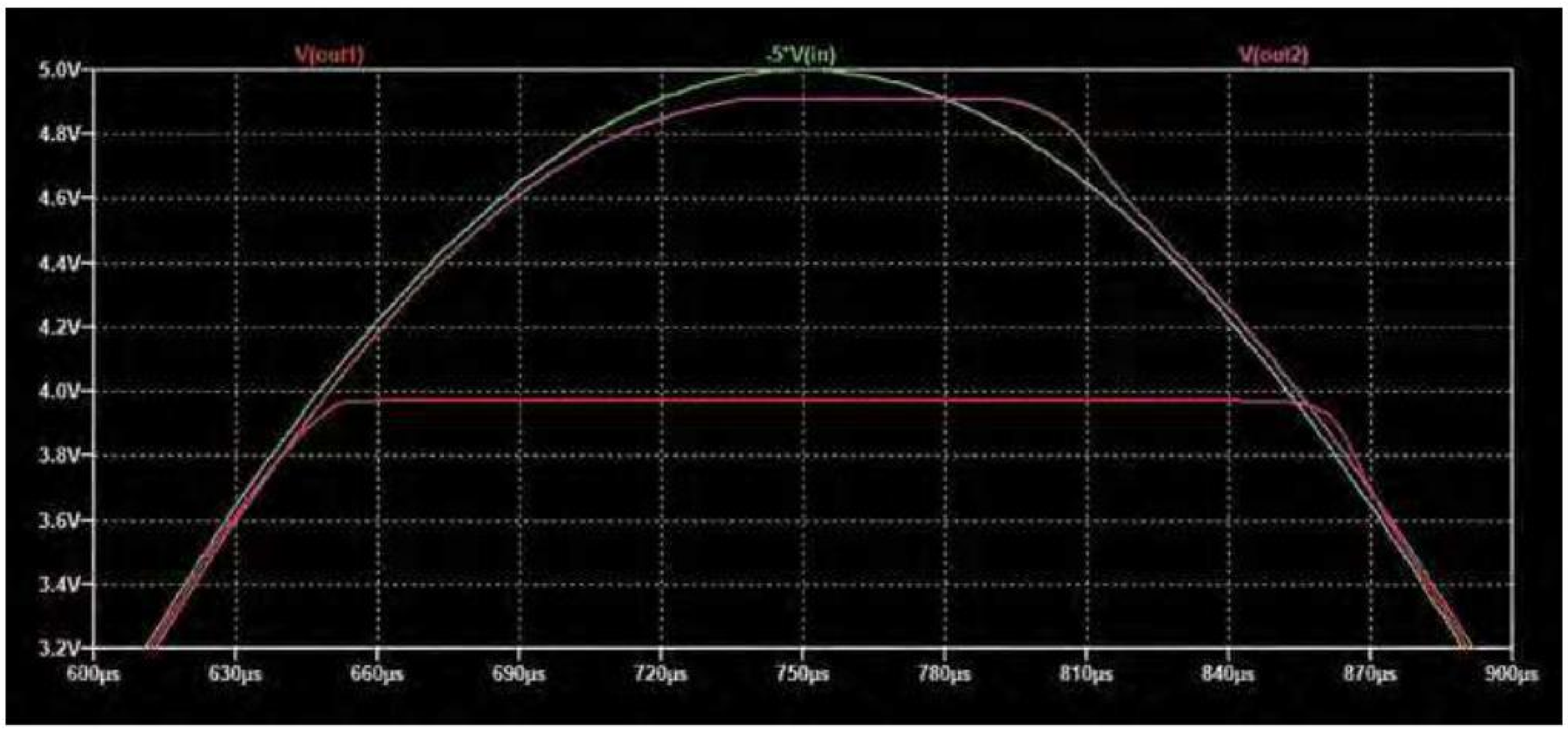


Fig.4. The same simulation as in Fig.1 but with Vin=1.0V. Compare with Fig.2, the rail-to-rail device's output is within 100mV of the supply, but the waveform is distorted.

±15V supply with a 10kΩ load. Some op amps have built-in DC-DC converter circuits to internally generate higher voltages than the supply to the chip. Such devices can produce fully rail-to-rail outputs.

Fig.1 is an LTspice simulation schematic for generating some illustrative waveforms. The circuit has two inverting amplifiers with a gain of 5 driven from the same input and operating on a ±5V supply. One amplifier uses an LT1001 'Precision Op Amp' and the other an LT1366 'Precision Rail-to-Rail Input and Output Op Amp' (these are arbitrary representative of each type). With the input at 0.96V peak, the output should be 4.8V peak.

Fig.2 shows the response of the two circuits and an idealised output obtained by directly plotting $-5 \times v_{\text{in}}$. The rail-to-rail device successfully outputs the signal – the peak is 200mV from the supply, but the standard op amp clips the signal at over 1V below the supply. Fig.3 shows the effect of increasing output current. This is the same situation except with the load resistors (RL1 and RL2 in Fig.1) reduced to 500Ω. Here we see that both outputs limit at a lower voltage, resulting in a significant increase in distortion from the rail-to-rail op amp.

The rail-to-rail op amp can get closer to the supply than 200mV in the example circuit – but again, at the expense of distortion. This is shown in Fig.4, where the input has been increased to 1.0V (with the loads at the original 2.0kΩ). Ideally, the output should have a 5V peak – exactly at the supply. The rail-to-rail device clips the signal less than 100mV below the supply voltage, but the large distortion would be unacceptable in many situations.

The waveforms in Fig.2 to Fig.4 indicates that that a simple figure of maximum output voltage may be insufficient when considering usable output range. The distortion produced by rail-to-rail op amps increases significantly as output levels reach a few hundred millivolts from the supply. At moderate loads, a limit of around 0.5V below the supply should avoid excessive distortion as a rule of thumb for many rail-to-rail devices, but it depends on output current and should be checked carefully if low distortion is important.

Rail-to-rail inputs

Op amps have differential inputs and amplify the voltage difference between their two inputs (the inverting and non-inverting inputs). In last month's *Circuit Surgery*, we discussed the basic BJT differential amplifier – this circuit forms the basis of the first stage of BJT op amps, although there are many refinements and variations in commercial op amp designs. As discussed last month, when dealing with differential signals and amplifiers we must also consider the common-mode input voltage. Given the two input voltages are V_1 and V_2, the differential signal (which is amplified) is $V_1 - V_2$ and the common-mode voltage is the average voltage at the inputs $(V_1 + V_2)/2$.

Op amps have very high gain and therefore in normal operation, in circuits such as amplifiers and filters, their differential input voltage is very small. Last month, we saw that the differential amplifier only provides linear amplification for differential inputs up to a few tens of millivolts. Large differential inputs of a few volts may damage the op amp by causing the base-emitter junctions of the input transistors to go into reverse breakdown (like a Zener diode).

Some op amps have maximum differential input voltages, around 600-700mV (or multiples thereof) due to the use of protection diodes to limit differential input voltage. Damage can still occur if currents through the protection diodes are not limited. Although exceeding a maximum differential input voltage of 600-700mV may seem difficult to avoid, it is actually unlikely to occur in standard op amp circuits with negative feedback

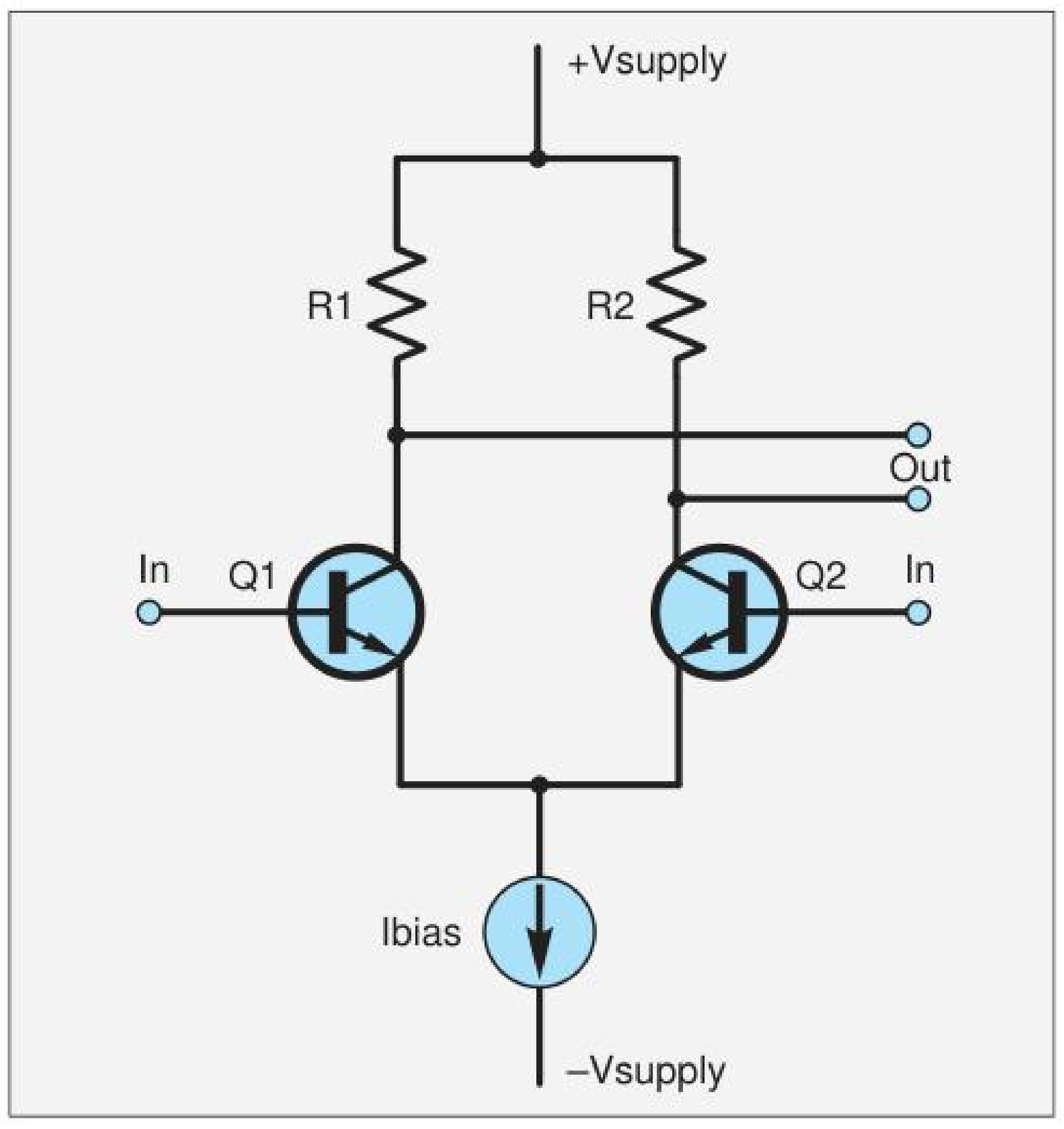


Fig.5. Differential amplifier created with a pair of NPN BJTs.

and it is often not an issue. However, some op amps have built-in resistors to limit the current and can withstand much larger differential inputs (eg, ±30V).

The maximum differential input is not directly related to the supply voltage. When op amps are described as 'having rail-to-rail input capability' this refers to the common-mode input signal. As we saw last month, the differential amplifier is a symmetrical circuit whose 'balance', and hence differential output, is not affected by changing the common-mode input voltage. However, this assumes a common-mode voltage somewhere in the middle of the supply range. If the common-mode voltage gets close to the supplies the operation of the circuit may fundamentally change and the differential amplifier either stops working or delivers much reduced performance.

Last month, we looked at the basic differential amplifier, as shown in Fig.5. In the context of this month's discussion it is worth considering the common-mode input voltage range over which it will operate. In order for the input transistors to be conducting they need a base-emitter voltage (V_{BE}) in the 0.6 to 0.7V range. If we assume the current source is the basic one we looked at last month, then there is a single output transistor, whose collector-emitter voltage must be above the saturation voltage (V_{CEsat}). The minimum voltage above the negative supply is the sum of these two voltages, so it is typically around 1V. If more complex circuits, such as higher-performance current sources are used, the 'stack' of transistors between the negative supply and input may be larger and have a bigger minimum voltage drop.

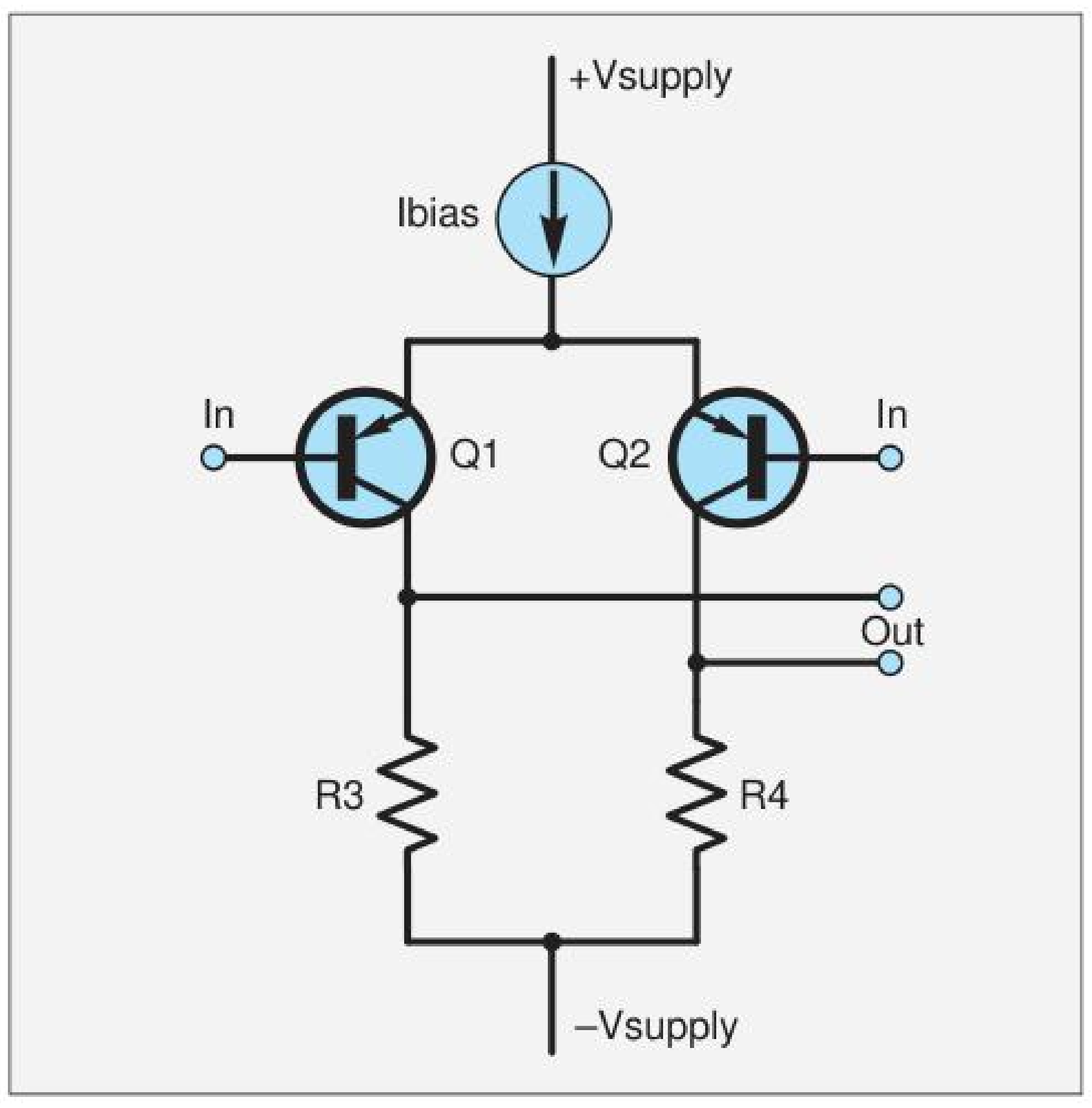


Fig.6. Differential amplifier created with a pair of PNP BJTs.

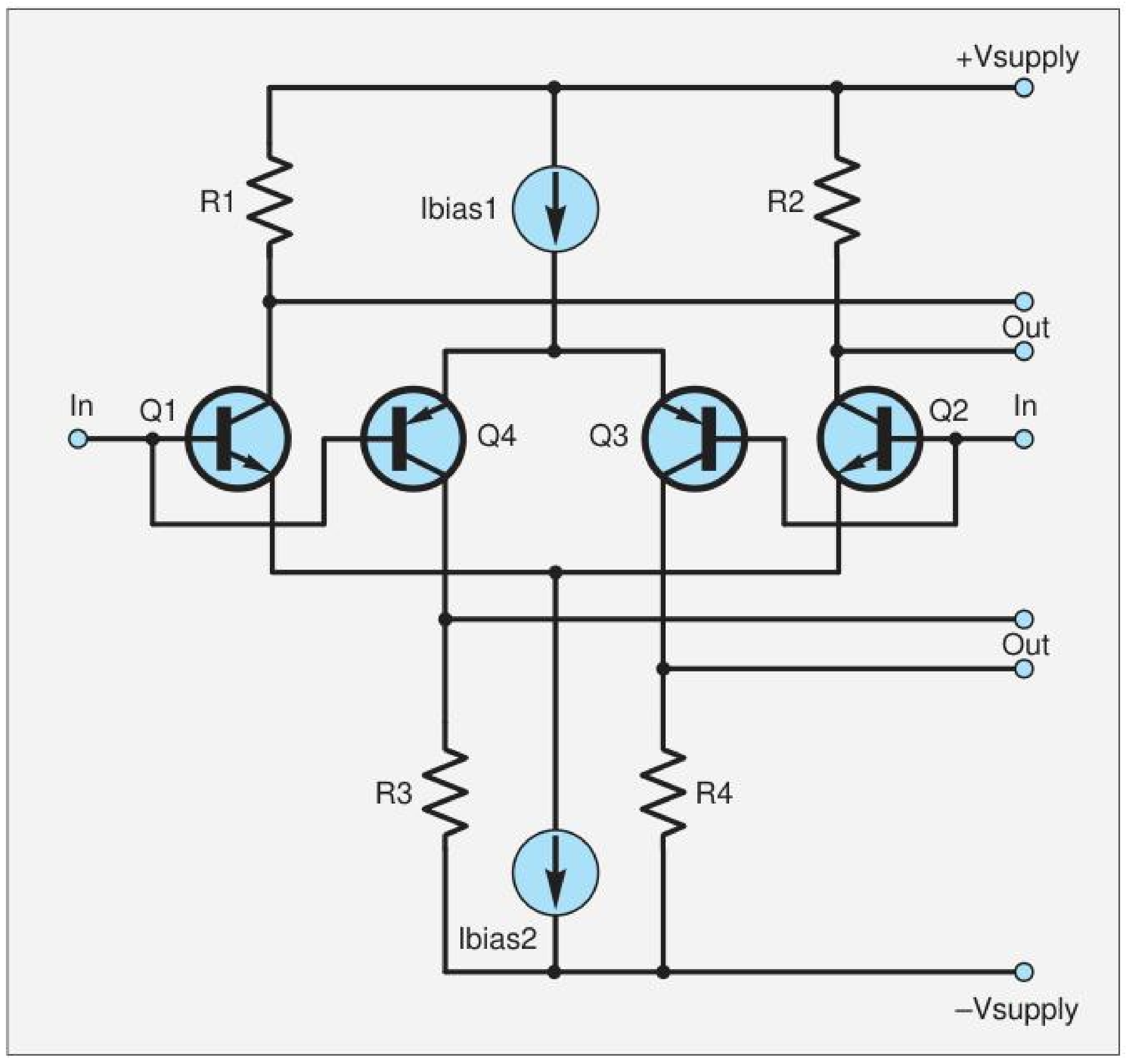


Fig.7. Rail-to-rail differential amplifier input stage – note the use of both NPN and PNP BJTs.

To find the maximum common-mode input voltage (with no differential signal) we have to find the condition for the input transistors going into saturation (we need their collector-emitter voltages to be larger than V_{CEsat} for acceptable circuit performance). Working through the circuit from the input to Q1, its emitter is at $V_{in} - V_{BE}$. The collector must be at least V_{CEsat} above this at $V_{in} - V_{BE} + V_{CEsat}$, and the supply is at the voltage dropped by R_1 ($I_{bias}R_1$) above the collector voltage. The supply is fixed, so we can write:

$$V_{Supply} = V_{in} - V_{BE} + V_{CEsat} + I_{bias}R_1$$

Rearranging this we can find the maximum V_{in} is given by:

$$V_{in} = V_{Supply} + V_{BE} - V_{CEsat} - I_{bias}R_1$$

Using a suitable choice of bias and R_1, the voltage drop across R_1 can be in (say) the 0.2 to 0.3V range, which with V_{BE} = 0.7V, and V_{CEsat} also in the 0.2 to 0.3V range, means that V_{in} can be around 0.1 to 0.3V above the supply voltage before the transistor saturates. So, this circuit can operate with common-mode voltages up to and just beyond the upper supply rail, but not all the way to the lower rail.

The same differential amplifier circuit can be implemented using PNP transistors, as shown in Fig.6. This reverses the relationship between the supplies and input – the input can go to, or beyond, the lower rail, but not to the upper rail. This makes the PNP circuit more suitable for single-supply op amps, where operation with the input at the lower rail (at 0V) is often needed – more on these circuits shortly. Rail-to-rail input op amps can be implemented using both NPN and PNP differential amplifiers in the same circuit – together they cover the entire supply range. The basic idea is shown in Fig.7, but this is a simplification. Circuits are needed to combine the output signals for the differential amplifiers. The op amp's characteristics (eg, offset voltage) may change with common-mode input voltage depending on which differential amplifier is active.

Single-supply op amps

Voltages are measured with respect to ground (0V) and often one of the supply rails will be ground – this is particularly likely in digital circuits. However, analogue signals are often bipolar – they can take on both negative and positive values – which means that if one of the supplies is ground then part of the signal (typically the negative half with a positive supply) will be outside the supply range. For this reason, it is common for analogue signal-processing circuits to have a split power supply; that is, two supply rails of equal and opposite voltages (for example +5V and – 5V, as in Fig.1). With split supplies, the signal, which is varying around 0V, is at a voltage which is in the middle of the supply range.

Split supplies provide a more straightforward design scenario for the internal circuitry of op amps, but single supplies have significant advantages in terms of size and complexity of the power circuits. Along with reduction in supply voltages over the years, there has been

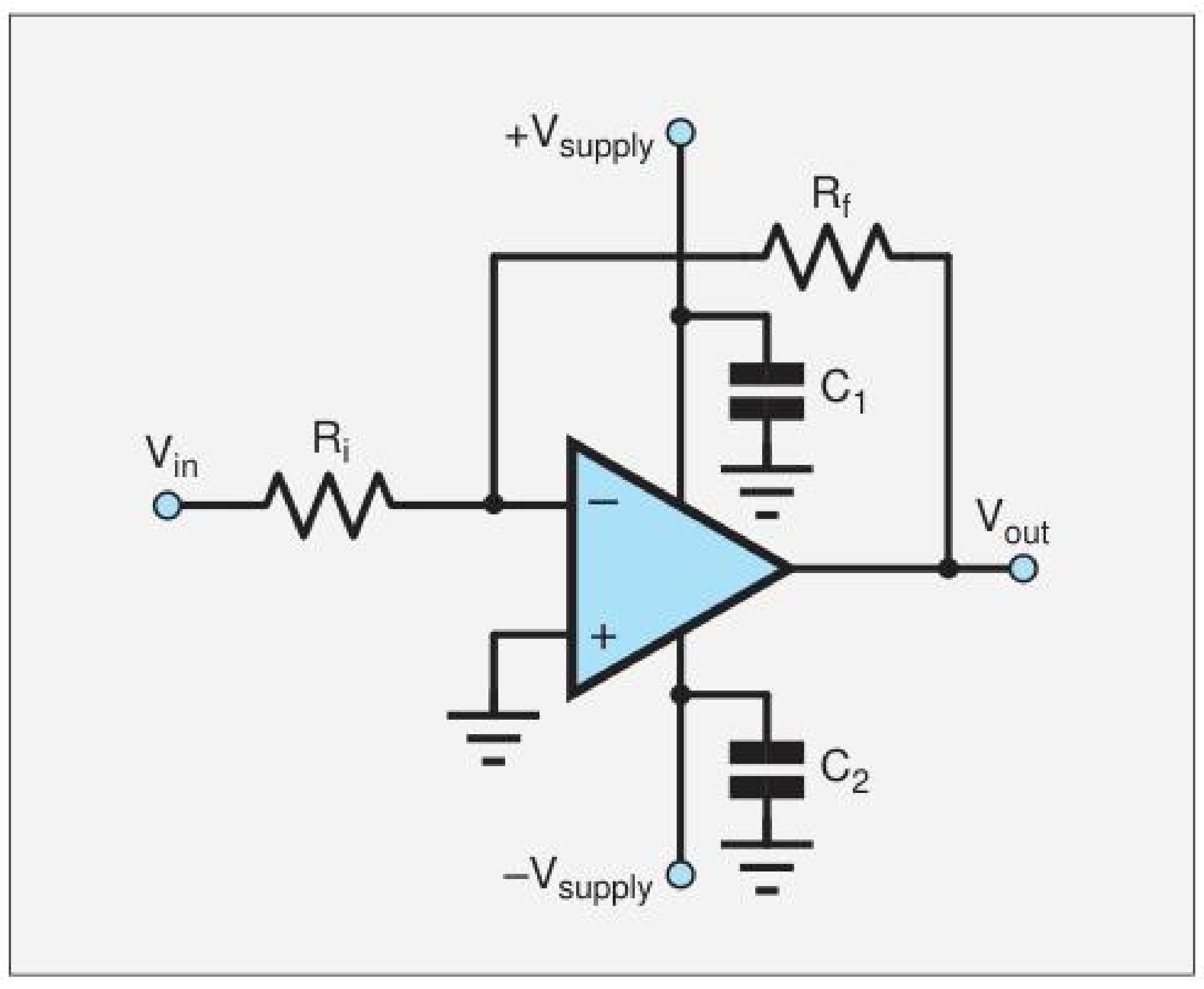


Fig.8. Op amp inverting amplifier with a split supply.

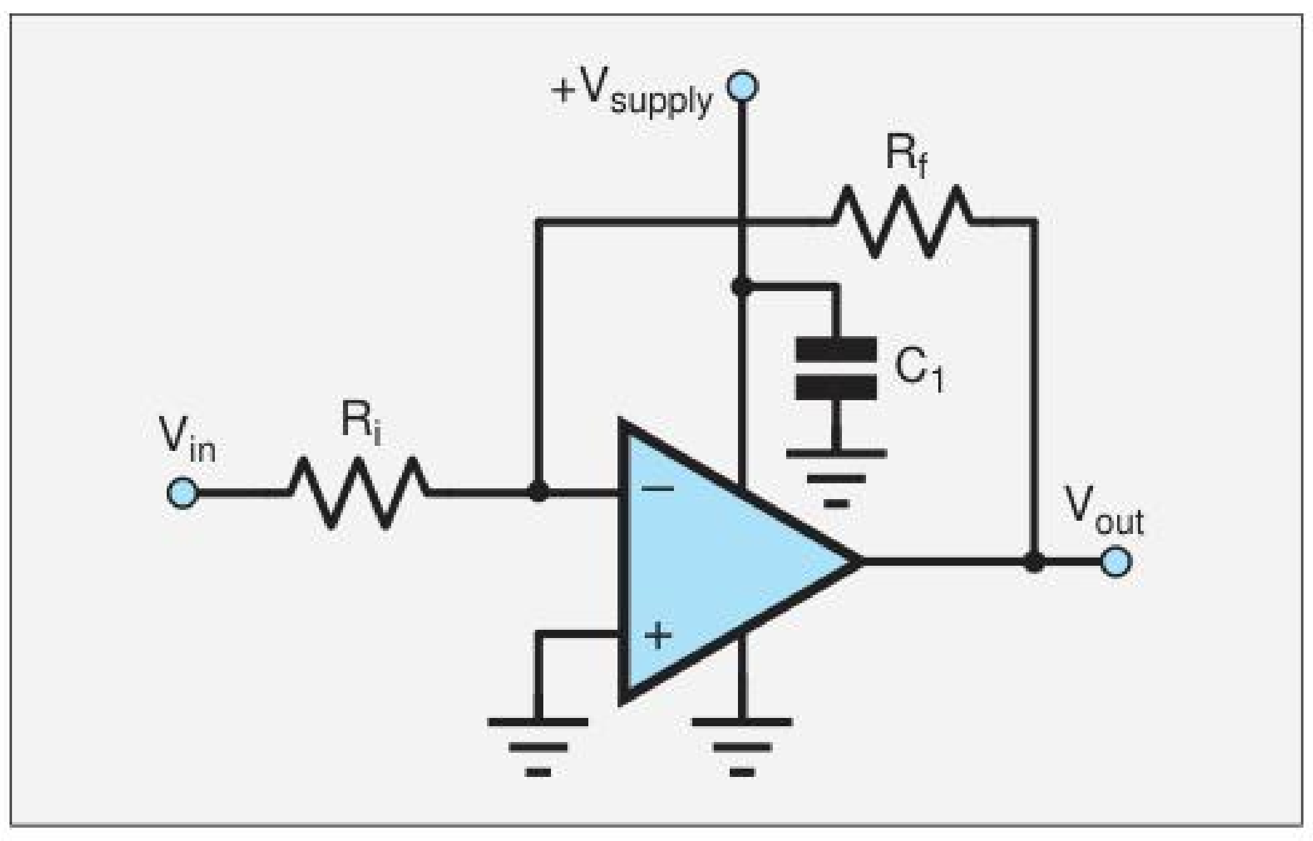


Fig.9. Op amp inverting amplifier with single supply.

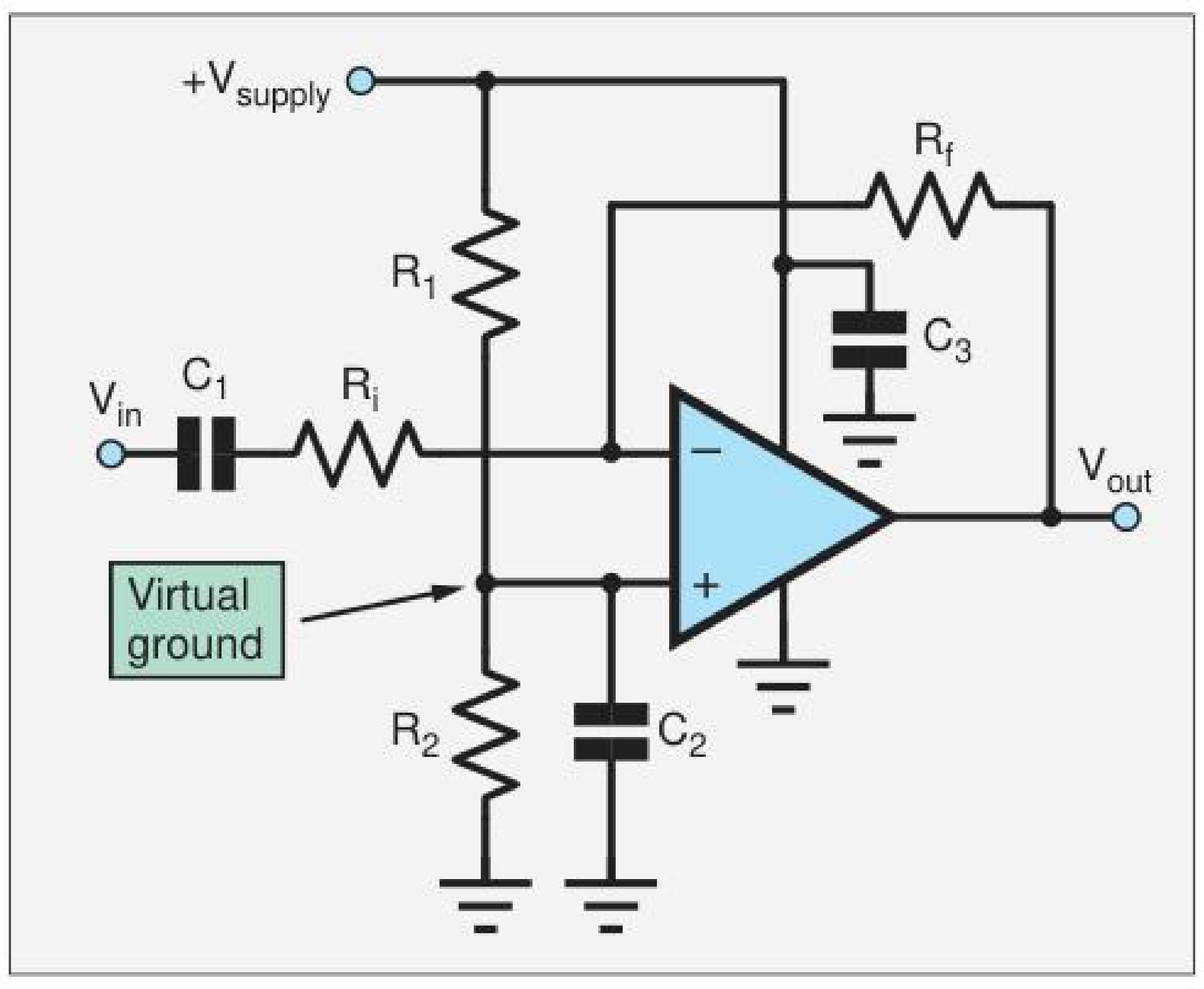


Fig.10. Single-supply op amp inverting amplifier with pseudo/virtual ground and capacitive coupling.

an increased use of single supplies. For circuits with a single supply it is not uncommon for it to be necessary for the circuit to operate with the input and/or output at 0V – this is equal to one of the supplies, so rail-to-rail type circuitry may be required for correct operation (at least for the ground side). It follows that not all op amps are suited to use with single supplies – at least if the applications are not limited to those only handling mid-range voltages. For this reason, op amp manufactures often make a point of stating when devices are suitable for single-supply operation.

Fig.8 and Fig.9 show the same op amp circuit – a standard inverting amplifier of gain $-R_f/R_i$ with two different power supply arrangements. Note the supply decoupling capacitors – you should always consult device datasheets for the specifics of what to use. Fig.8 shows a split supply with voltages of $\pm V_{Supply}$ and Fig.9 shows a single supply of ground (0V) and $+V_{Supply}$. With $R_i = 1k\Omega$ and $R_f = 5k\Omega$ the gain is 5 (as in Alexandru's example and Fig.1). If we have an input signal which is a 0.5V peak sinewave centred on 0V then the circuit in Fig.8 will happily output a 0.5 × 5 = 2.5V peak sinewave, also centred on 0V. Unfortunately, the circuit in Fig.9 will not work because the amplifier is inverting, so a positive input signal should result in a negative output voltage – which is outside the supply range and not possible. Even a rail-to-rail op amp will not help here as the output would be required to go well outside the supply range.

Op amps with negative feedback, such as the circuit in Fig.9, control their output voltage such that the input voltage difference is as close to zero as possible. Their very high gain means that the input voltage difference is very small and a 'close to zero' input difference is achieved. With negative feedback active, the op amp's inputs behave almost like they are shorted together (a virtual short circuit). For the circuit in Fig.9 the non-inverting input is wired to ground, so the inverting input will also be at 0V if the op amp is acting as a linear amplifier.

In general, it is the signal at the inputs of the op amp itself – not of the whole circuit – which is what matters in terms of rail-to-rail capability. As discussed above, this will be very small, so if the op amp common-mode voltage is within the usable range the signal should not cause a problem. Therefore, if the op amp has rail-to-rail input capability, the circuit in Fig.9 may be able to operate correctly in terms of the input, but as already noted it will not be able to output the full waveform of an AC waveform centred on 0V. If the op-amp is not designed for single-supply use with common-mode input at the negative rail then the circuit will not work at all. The exact non-functional behaviour will vary with different op amp types.

Virtual ground

A solution to the problem with the circuit in Fig.9 is to create a virtual or pseudo ground at half the single-supply voltage and capacitively couple the input signal. This is shown in Fig.10, in which the virtual ground is produced by the two equal resistors R_1 and R_2 and the capacitor C_2. R_1 and R_2 typically have values in the tens to hundreds of kilohms range. Higher values reduce the drain on the supply and the op amp should not need to take large currents from the divider, so low values should not be needed. C_2 plays a similar role to a supply decoupling capacitor and reduces noise on the virtual ground. Virtual grounds like this can degrade performance and more sophisticated circuits can be used – for example, buffering the potential divider with a unity-gain op amp amplifier. Texas instruments make a 'rail splitter' precision virtual ground IC, the TLE2426.

C_1 in the circuit in Fig.10 removes the DC component from the input (typically 0V) and just allows the AC signal through to the op amp. Another effect of C_1 blocking DC is in the relationship between the virtual ground voltage and the output. Fig.11 shows the DC equivalent circuit for Fig.10 – this is obtained by replacing the capacitors with open circuits. With C_1 open, R_i is disconnected at one end, so is completely removed. For Fig.11, we see that as far as DC is concerned the op amp has 100% negative feedback. The circuit effectively has the virtual ground connected to a unity-gain buffer. Thus, the DC output of the op amp is 1× the virtual ground voltage. Using equal resistors (as indicated above) will give half the supply as both the virtual ground and the output with no signal present. If an output centred on 0V is required, then an output coupling capacitor can be used.

AC coupling the input makes the virtual ground straightforward to use. If we don't use C_1 in Fig.10 then we can amplify a DC input, but the virtual ground setup is more complex. A DC equivalent circuit of Fig.10 with C_1 removed and the input connected to a source with 0V DC under no signal conditions is shown in Fig.12. The circuit is now a non-inverting amplifier as far as the virtual ground is concerned. With an inverting gain of –5 the non-inverting gain is 6, so the virtual ground needs to be at 417mV to give 2.5V DC at the output with no signal – hence different resistor values are needed in the potential divider. This illustrates the fact that single-supply op amp circuits can be more difficult to work with than traditional split-supply versions.

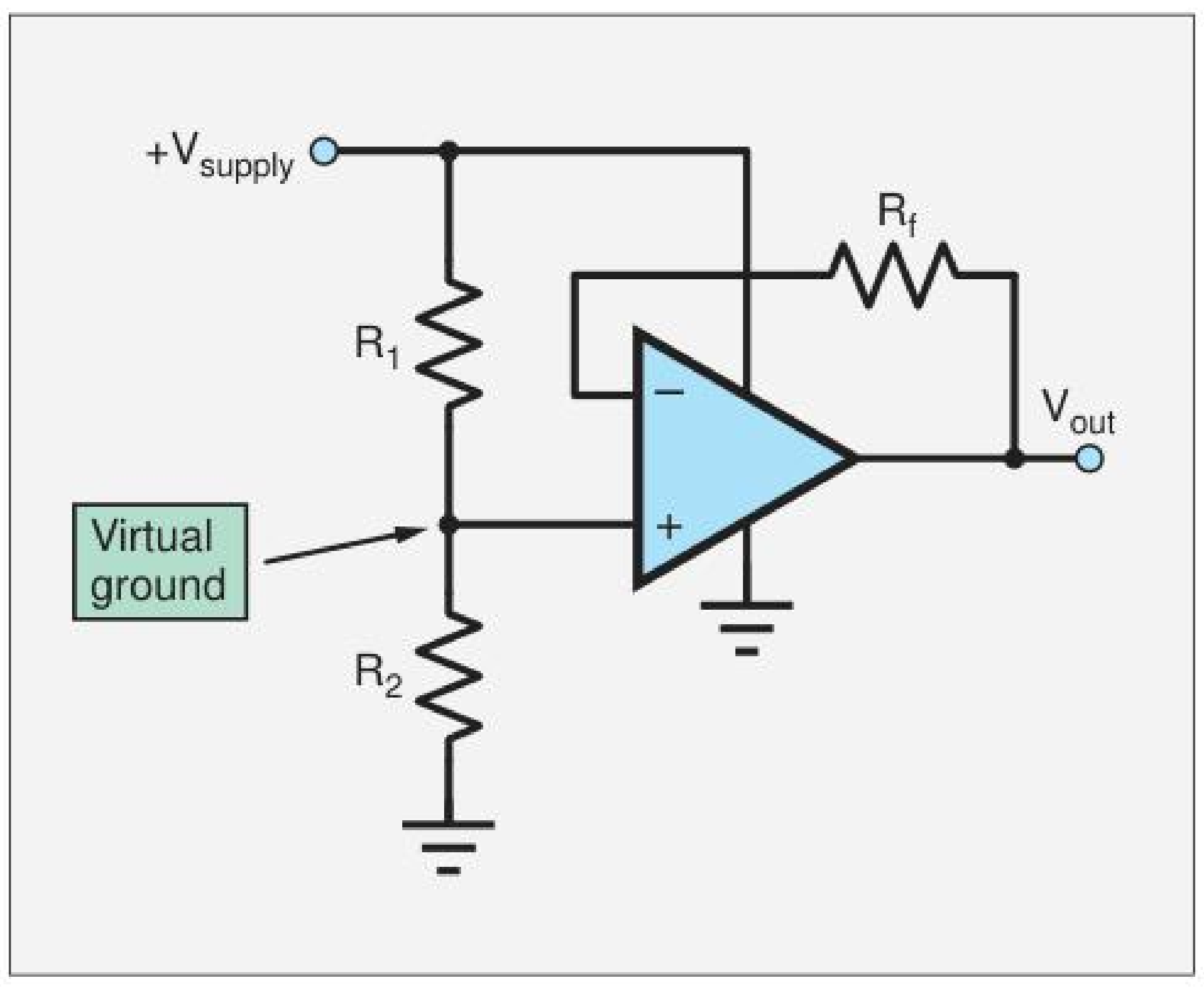


Fig.11. DC equivalent circuit for Fig.10.

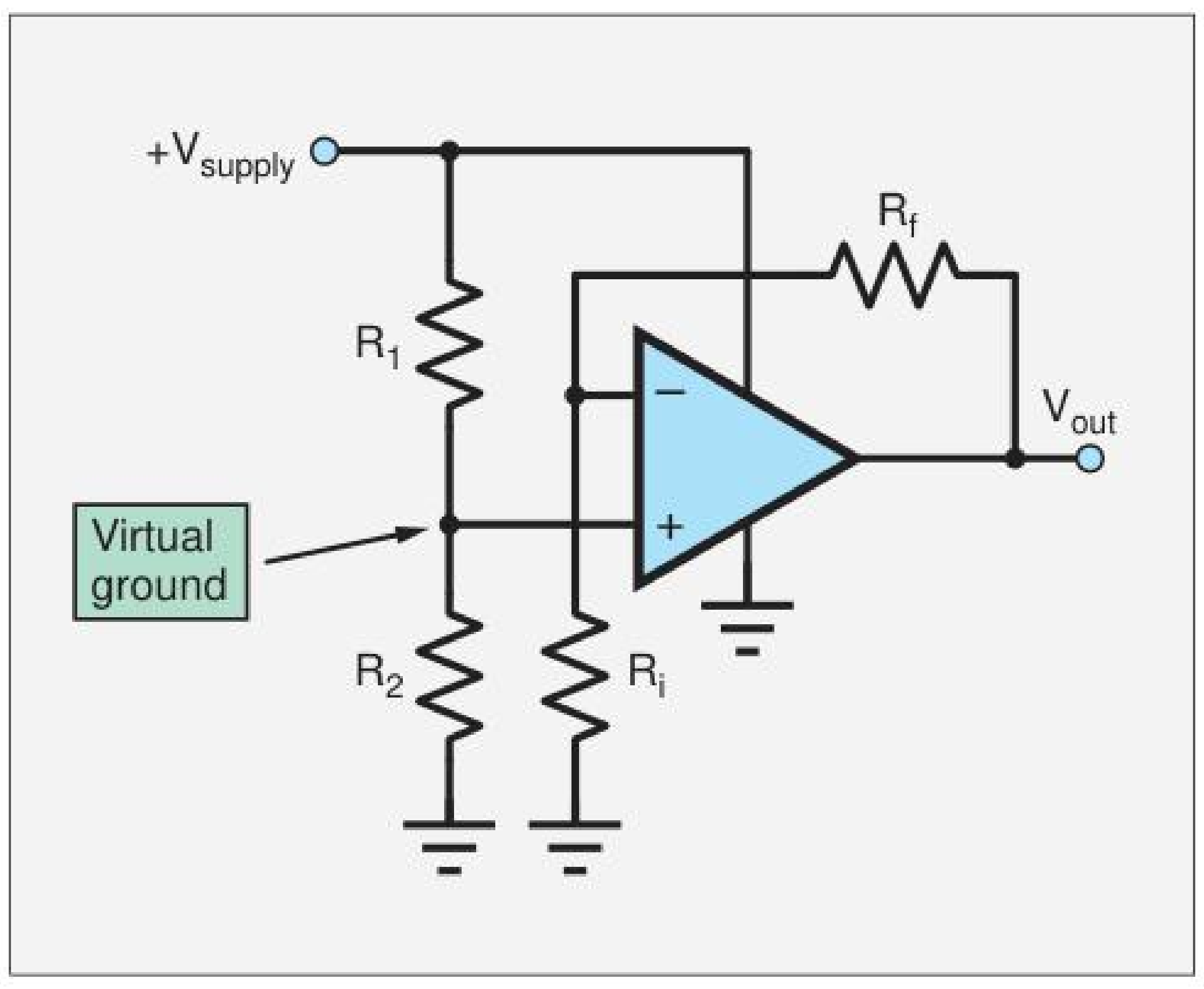


Fig.12. DC equivalent circuit for Fig.10 with C_1 removed (DC input)

Simulation files

Most, but not every month, LTSpice is used to support descriptions and analysis in *Circuit Surgery*.
The examples and files are available for download from the *PE* website.

Practically Speaking

Hands-on techniques for turning ideas into projects – by Mike Hibbett

Introducing the K40 laser cutter and engraver

Hard to believe, but it's been 13 years since we first discussed the beginning of affordable hobbyist 3D printers. This month, we explore another tool that has moved from high-tech industrial application to the hobbyist space – the laser cutter.

Industrial laser cutters come in many sizes, from relatively small, low-power (100W and under) used for engraving, up to large 1800mm-square machines used for cutting clothing designs, and much larger 15kW CNC machine tools for cutting steel plate (**https://youtu.be/tbSfc_a1PQg**). For our purposes, materials such as wood and plastics of up to 12mm thickness can be cut with relatively low power and relatively inexpensive lasers, in the 30W – 60W range, and it is this capability and pricing that has driven the availability of hobbyist-level machines.

The K40 laser cutter

One such machine is called the 'K40'. Unusually, it is a completely open design, so there is no single manufacturer that owns the copyright – yet the specification remains common (with minor differences) across dozens of different manufacturers. An example machine is shown in Fig.1. It consists of just a few modular components:

- *x-y* mechanical bed, similar to a 3D printer
- Controller circuit, also similar to a 3D-printer controller
- 22kV high-voltage power supply
- The laser.

The laser is nothing more than a long glass tube, much like a hugely oversized valve (US tube) from the 1950s. You can see the tube in our machine in Fig.2 – it is easily accessible from a hinged cover at the rear.

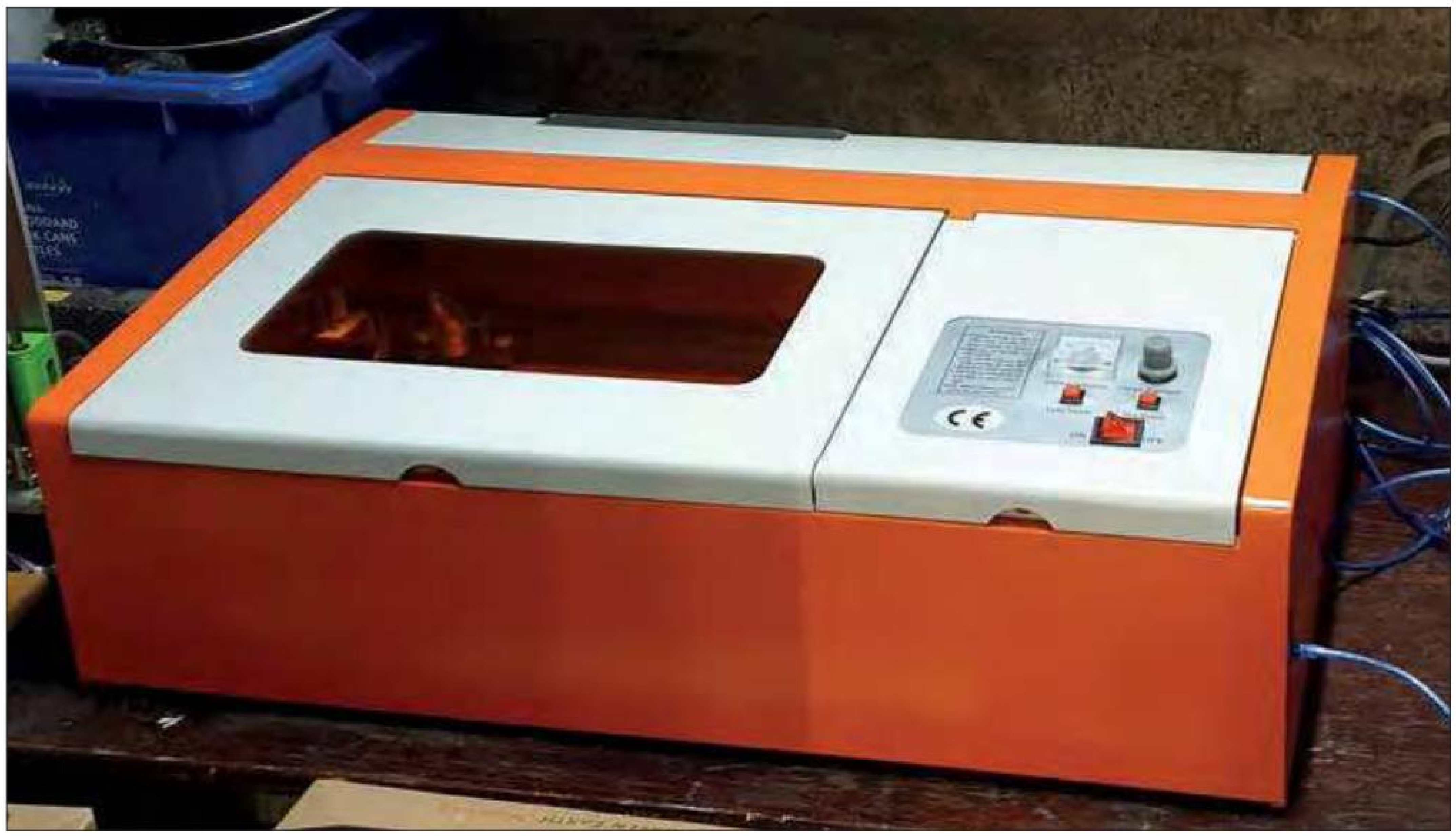


Fig.1. The popular, flexible and inexpensive K40 laser cutter and engraver.

WARNING!

We highlight several safety issues in this article:

- **Eye protection**
- **Fume extraction**
- **Wiring/earthing**

These are not nice-to-have optional extras – you must follow all safety guidelines to protect yourself and those around you.

Mirrors fitted to the enclosure and the *x-y* bed direct the laser beam through a small lens onto the work surface. The controller controls the position for the focusing lens in the *x* and *y* coordinates, and also turns the laser on and off, allowing very precise and repeatable cutting. Etching – as opposed to cutting – is achieved by manually reducing the power output of the laser through a simple control.

There are two variations on the K40: one with an analogue control and display of the laser power, the other with a simple digital display of the percentage of maximum power output.

The analogue display is considered by the community at large to be the better choice as it is more meaningful – knowing the current flowing through the laser tube allows you to get a better picture of how hard you are driving it. Minimising the power, thereby avoiding over-driving the tube, can significantly prolong its life. A laser tube is considered a consumable, with an operational life of approximately 1000 hours. Replacement tubes cost around £70 – £100. You can

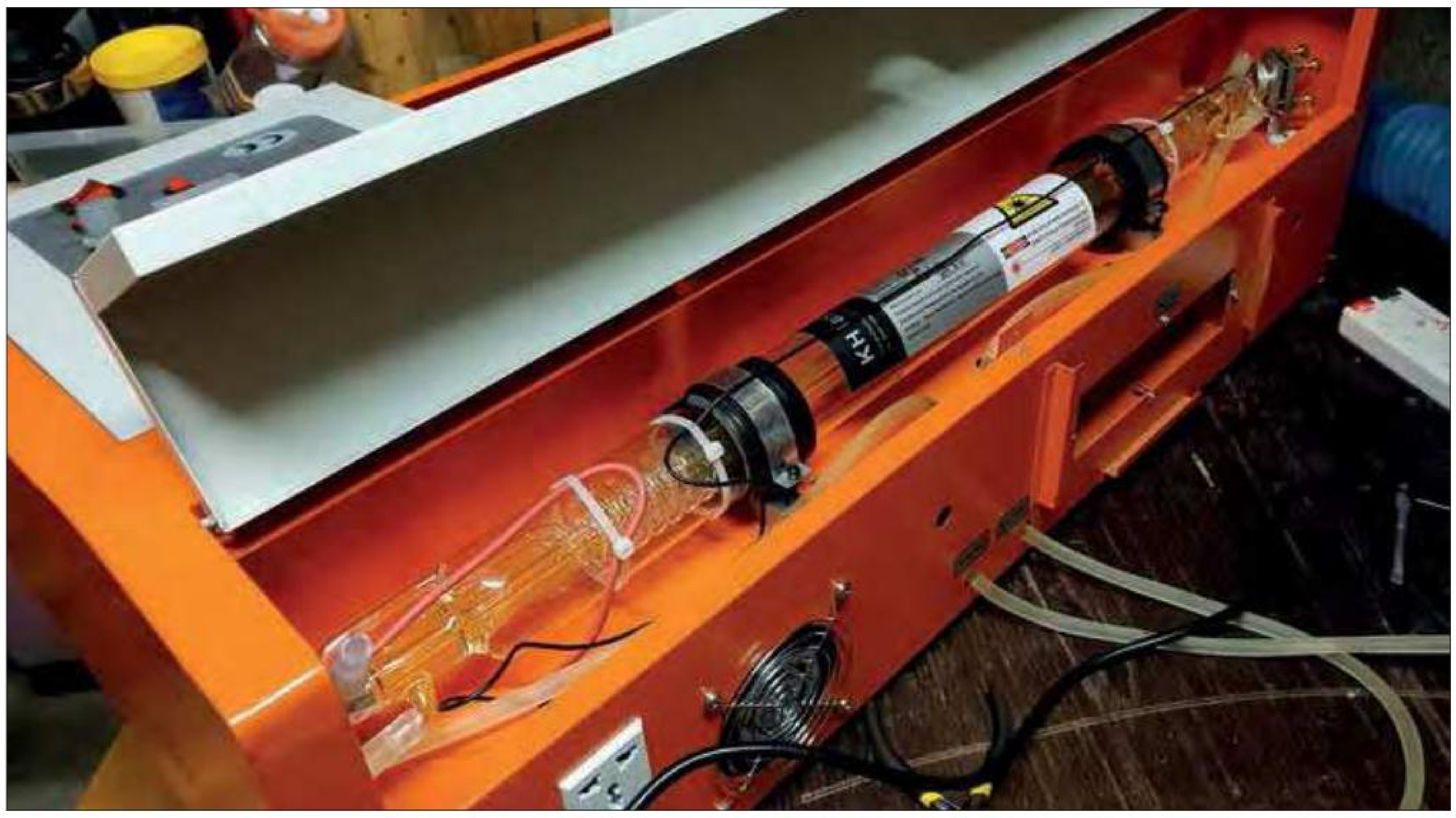


Fig.2. The K40 laser tube – it's delicate, so check it for damage as soon as it arrives.

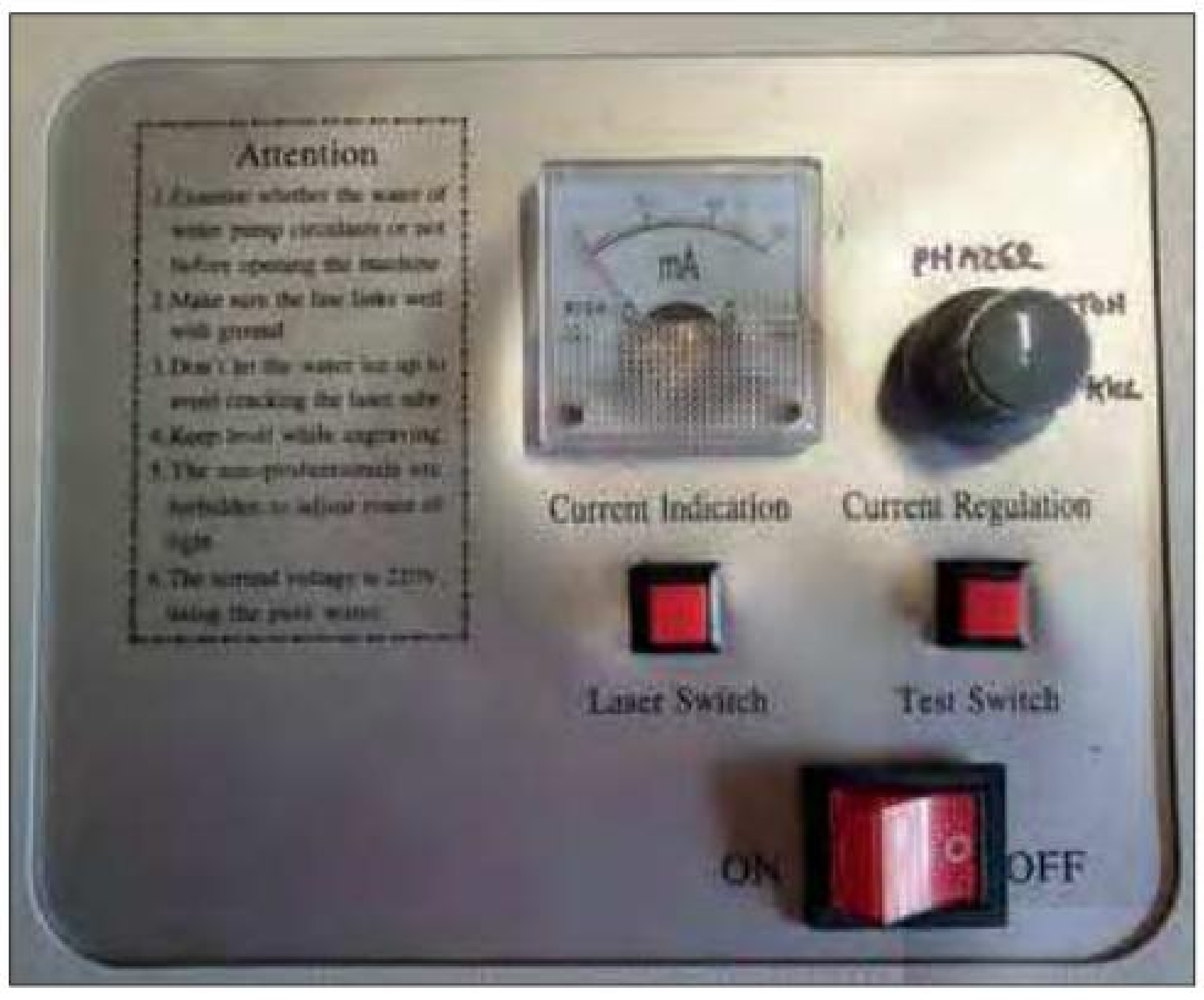


Fig.3. Analogue K40 control panel.

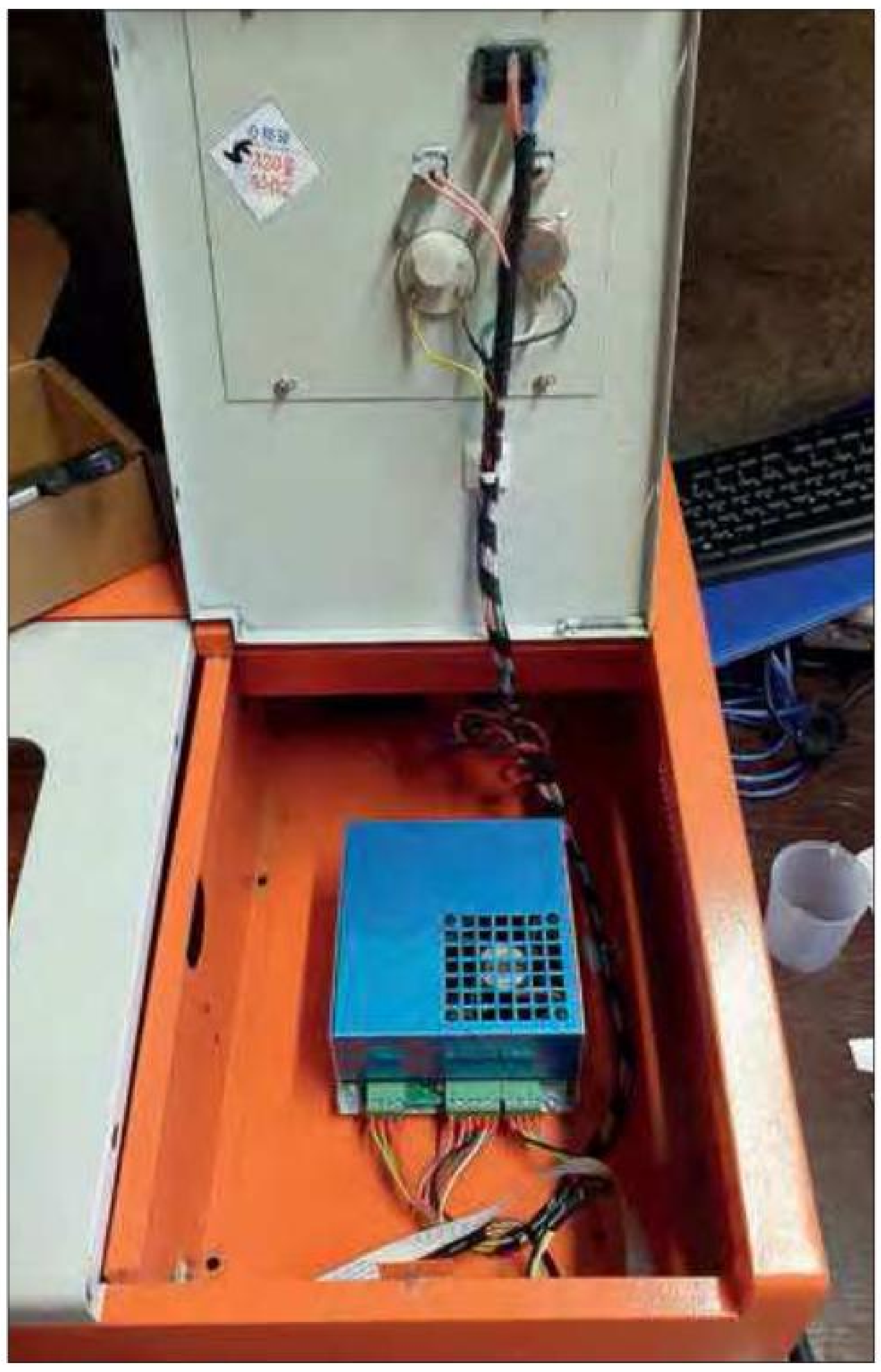


Fig.4. Inside the K40 electronics area.

see an example of an analogue control panel in Fig.3.

With many suppliers to choose from and prices ranging from £300 – £600, we went with a machine priced at £333 from an Amazon UK supplier based in Germany. Slightly cheaper deals are available through Far East suppliers, but we wanted to avoid the risk of attracting import duties. Also, our seller sold many other etching and laser-cutting tools, so seemed to be a specialist, and might be knowledgeable in the appropriate handling and shipping of the technology.

What is supplied?

The machine arrived in a large cardboard box, weighing in at 20kg and measuring approximately 900mm × 600mm × 375mm. It took two people to move it into the house and then to the workshop.

To our relief, there were no visible signs of damage; it's not unusual for these machines to be delivered with dents, or even broken laser tubes. We were prepared for some issues, but none were found. Perhaps the seller, based in Germany although clearly fronting a Chinese organisation, did some quality assurance first… perhaps we got lucky.

Inside the box is the machine itself, a water pump and flexible tubing (the laser tube is water cooled) an air extraction fan, and a length of very cheap 100mm-diameter extendible ducting to allow venting of the toxic fumes from the device to, well, somewhere else.

A somewhat simplistic user-guide and control software on a USB stick were the final contents. Following the advice from the Internet these were discarded; there are far better and (frankly) more honest machine setup tutorials on YouTube, and far better control software available for free.

Extras needed

The machine must be water cooled, and for this you'll need at least a bucket to hold about a gallon of de-ionised water – available from DIY stores. Anti-freeze should be added too if you live in a cold climate.

Two mains plugs are recommended to replace the Chinese ones fitted to the pump and extractor fan. To make life simple, a switched mains socket strip for all three is advised.

Unless you like to gamble with your eyesight, a pair of 10600nm wavelength protective goggles are a must. You can expect to pay £40 for a reasonable pair. Given that the laser beam is invisible, and that misalignment of the mirrors during shipment is not uncommon, this is a safety precaution you must take. We borrowed a pair from the local hacker space while checking the alignment.

You'll also need a PC to control the machine – we had an old laptop that found a new lease of life. The control software requirements do not require a powerful machine – the K40 can even be operated with a Raspberry Pi at a push. Control is through a standard USB cable connection.

Installation

Once unpacked, the machine measures 84cm wide, 70cm deep and 27cm tall – but you will need 67cm clearance above any table it sits on to open the lid. You will also need somewhere to place the bucket of water.

Fume extraction posed a greater challenge; in the end, we moved our machine closer to an outside wall, and drilled a 100mm-diameter hole, followed by installing standard air vent plates to the outside wall. Adequate venting is essential, as the gasses given off during cutting vary from unpleasant (if cutting or etching wood) to highly toxic (if cutting unspecified plastic.)

We removed the two Chinese power plugs and fitted UK mains plugs. Two mains outlets are provided on the rear of the machine, but we found them loose, and preferred the mechanical stability of the larger UK plugs.

The internals of the machine are easily accessed and partitioned simply – the laser tube area is at the rear, stretching the entire width of the machine; and the electronics area (with plenty of room for additions) is front right, as shown in Fig.4. The work area covers the rest of the front, accessible via a hinged lid, as seen in Fig.5. There you can see two of the mirrors that direct the (invisible) laser beam.

Submerging the no-name, mains-powered water pump into, well, water, was a difficult challenge. We ended up putting it in a large sealed plastic container and labelling it 'Don't touch when powered!'. There were no warnings from other users about the use of this pump on the Internet, but best to be safe. After eight months of use, however, we have had no trouble with it.

The air pump, on the other hand, was another matter. We were pre-warned on the Internet about the quality of wiring on the pump and were not disappointed – ours was wired exactly as discussed on

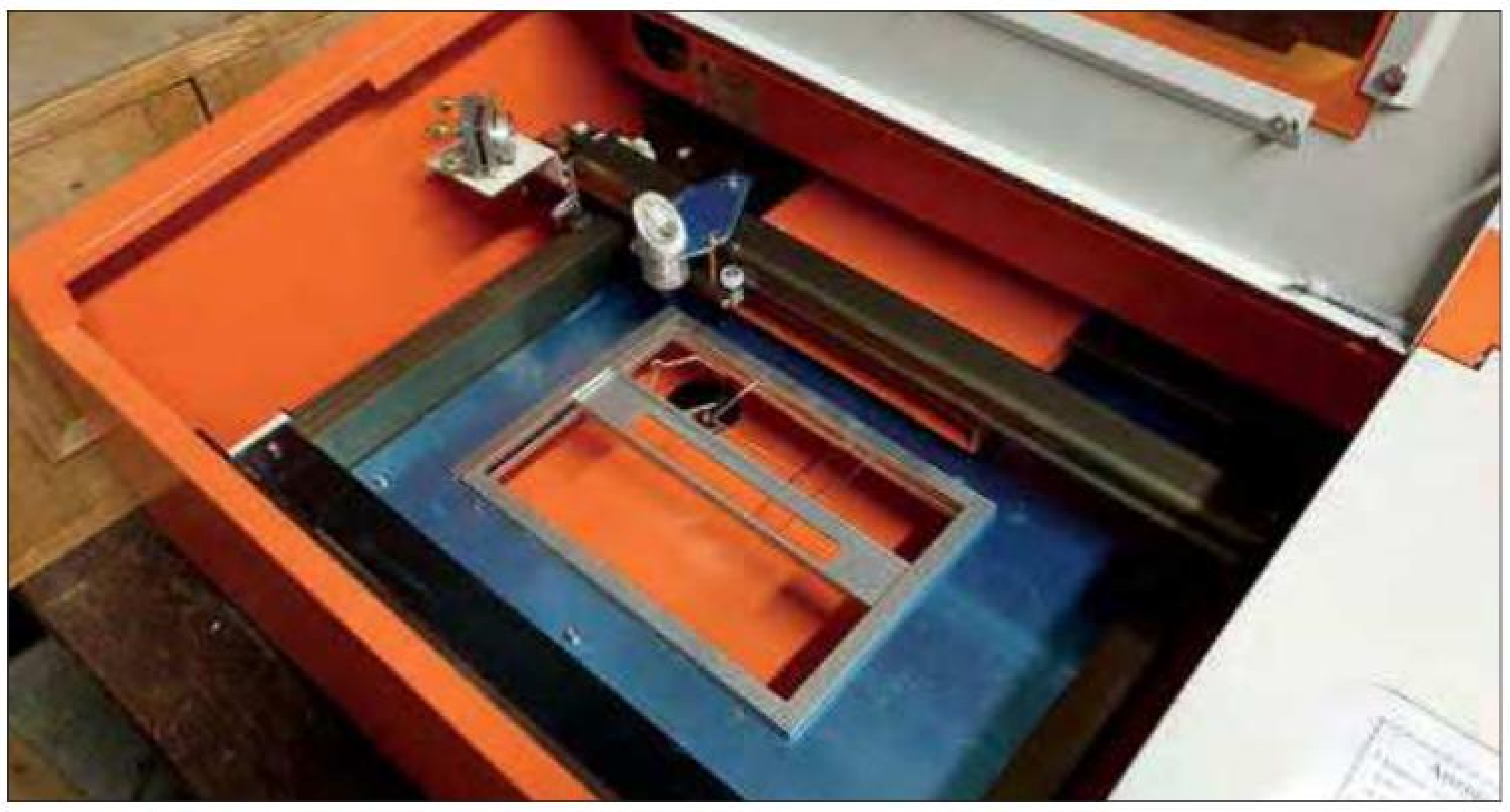


Fig.5. The K40 work area – it's easy to upgrade this area's size.

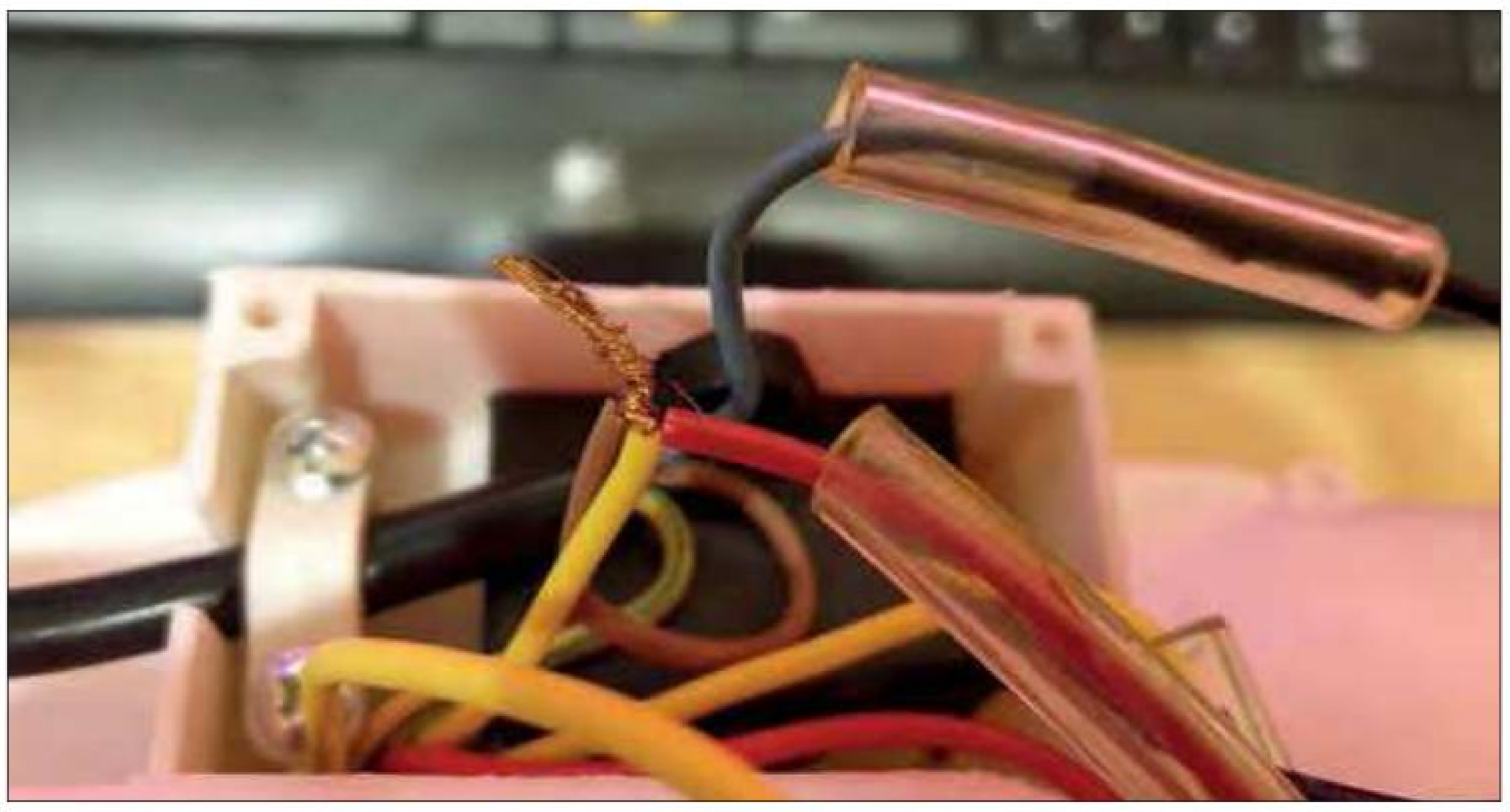


Fig.6. You must check the air pump wiring – see text!

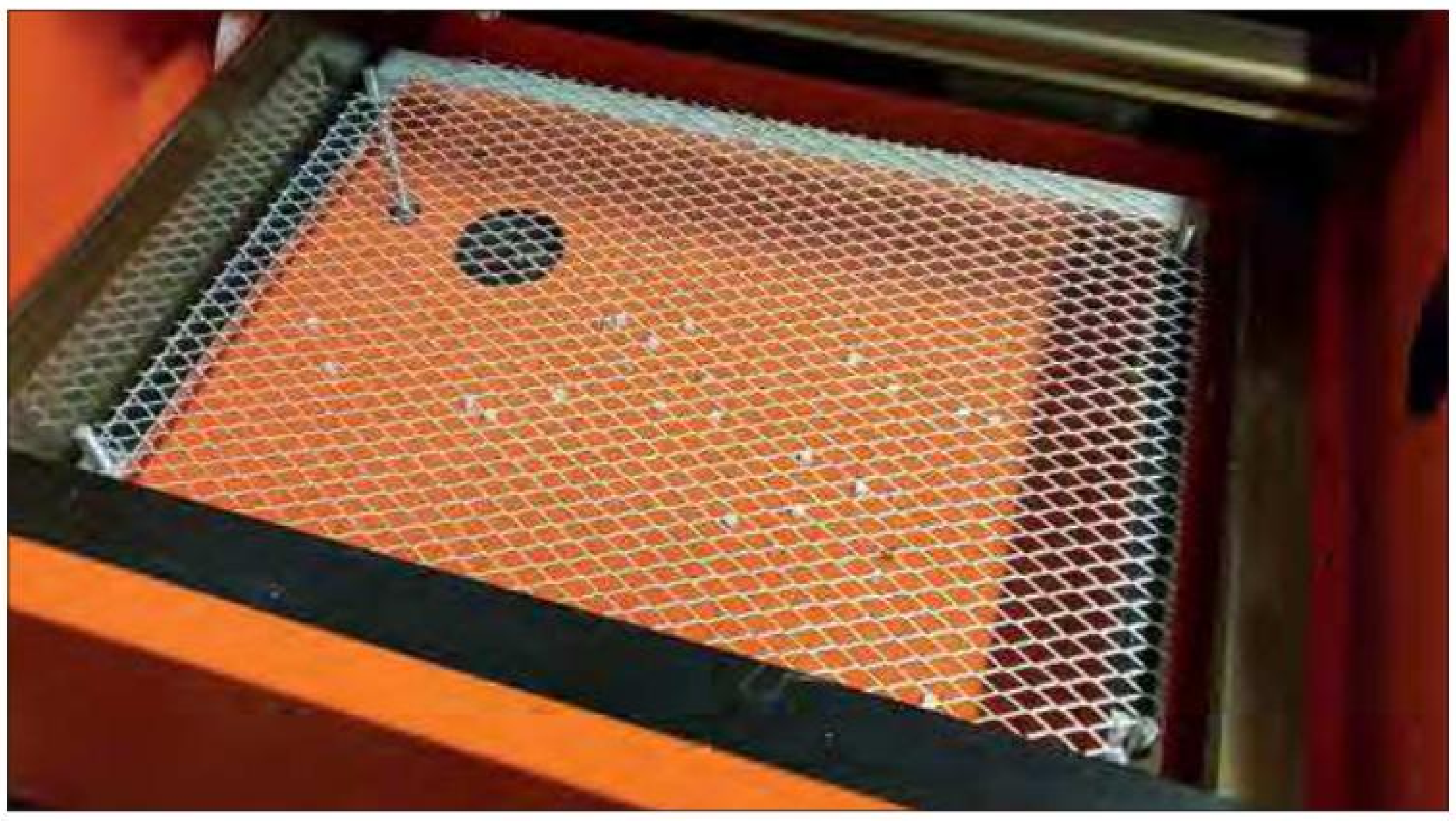


Fig.7. The upgraded K40 work area

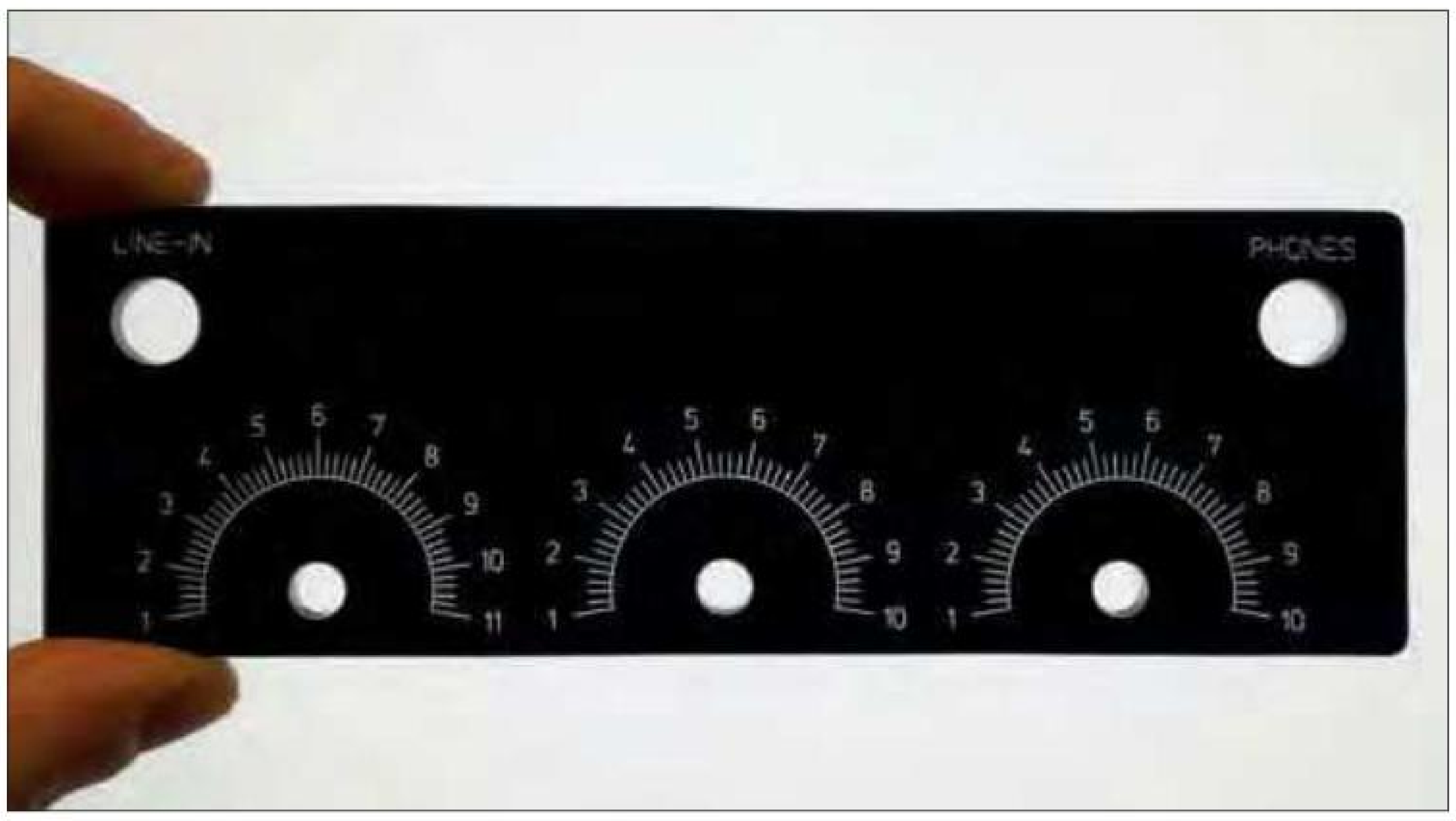


Fig.8. Very nice! – a backlit front panel with etched scales.

forums, as shown in Fig.6. The mains wires from the input supply are simply twisted to the motor wires! It's unbelievable that Amazon UK permit this equipment to be sold anywhere; the device's CE approval mark mean absolutely nothing in today's Internet-based consumer markets. Buyer beware! Thankfully, we had done our research and were forewarned of these horrors, so we soldered the wires and fitted insulation.

Another point we were pre-warned about was the fitting of the air extraction coupler. This (which is a combined coupler and fan unit) is very loose when fitted to the machine, and you would be well advised to fit at least duct-tape to hold it to the laser cover, or use some better solution (again, the Internet has many suggestions for improved air exhaust couplers, but they do require access to a 3D printer.) We were happy with the performance of this exhaust fan and so simply duct-taped it to the enclosure.

Toxic plastic fumes

This is a good time to mention fumes: when cutting plastic, perspex is safe, with appropriate venting. Plastics such as PVC emit *highly* toxic gases when burnt that will injure both you and the machine. You must research the material you are thinking of using before cutting or etching. Wood, paper, cardboard and perspex are fine, with appropriate venting. Other materials – assume they are deadly, until research says otherwise. All cutting operations give off bad-smelling smoke, so even venting out of an open window can be a problem, both in terms of back draught and sending the smoke into your home, or to your neighbour. We have two friends who own laser cutters but accept they cannot use them during the summer due to neighbour's open windows. Fortunately, our vent is into an unused alleyway.

The work surface comes with a simple spring-loaded clamp which restricts the work area to 21cm × 9cm, as seen in Fig.5. This is surprisingly small given the volume of the work area. Thanks to the very simple design of these machines, modification to accommodate a much larger working area is straightforward using just common DIY tools and £20 worth of materials. This is considered a simple yet essential upgrade. You can see our solution in Fig.7. This gives a much nicer 30cm × 20cm work area.

The controls

The machine controls are very simple amd consist of just the following three items:

- Potentiometer to adjust the power level, with an analogue dial showing the value
- Laser ON/OFF switch, which allows you to test control of the machine without firing the laser
- Momentary push button which allows you to manually turn the laser on. This is useful for checking your power level setting and confirming the exact position of the laser head (handy when you are trying to minimise the amount of waste material).

Software

As mentioned, you can ignore any software or user guides supplied with the machine; the K40 laser is very well supported on the Internet by a huge community of fellow makers. A free program has been developed specifically designed to control this printer, called 'K40 Whisperer'. It has become the standard control software of choice for hobbyists.

Unlike 3D printers, which require design files to be created in a 3D CAD program, laser-cutting profiles are two dimensional so they can be created with much-simpler 2D drawing tools such as Inkscape. Images must be in SVG format, but if you are looking to etch photos or drawings, Inkscape can convert them for you.

If you are interested in designing front panels, The US company Front Panel Express provide a free program, Front Panel Designer, which make this task easy. Our back illuminated amplifier panel, shown in Fig.8, was designed in just a few minutes, the file exported as an SVG image, then cut and etched in just a few minutes.

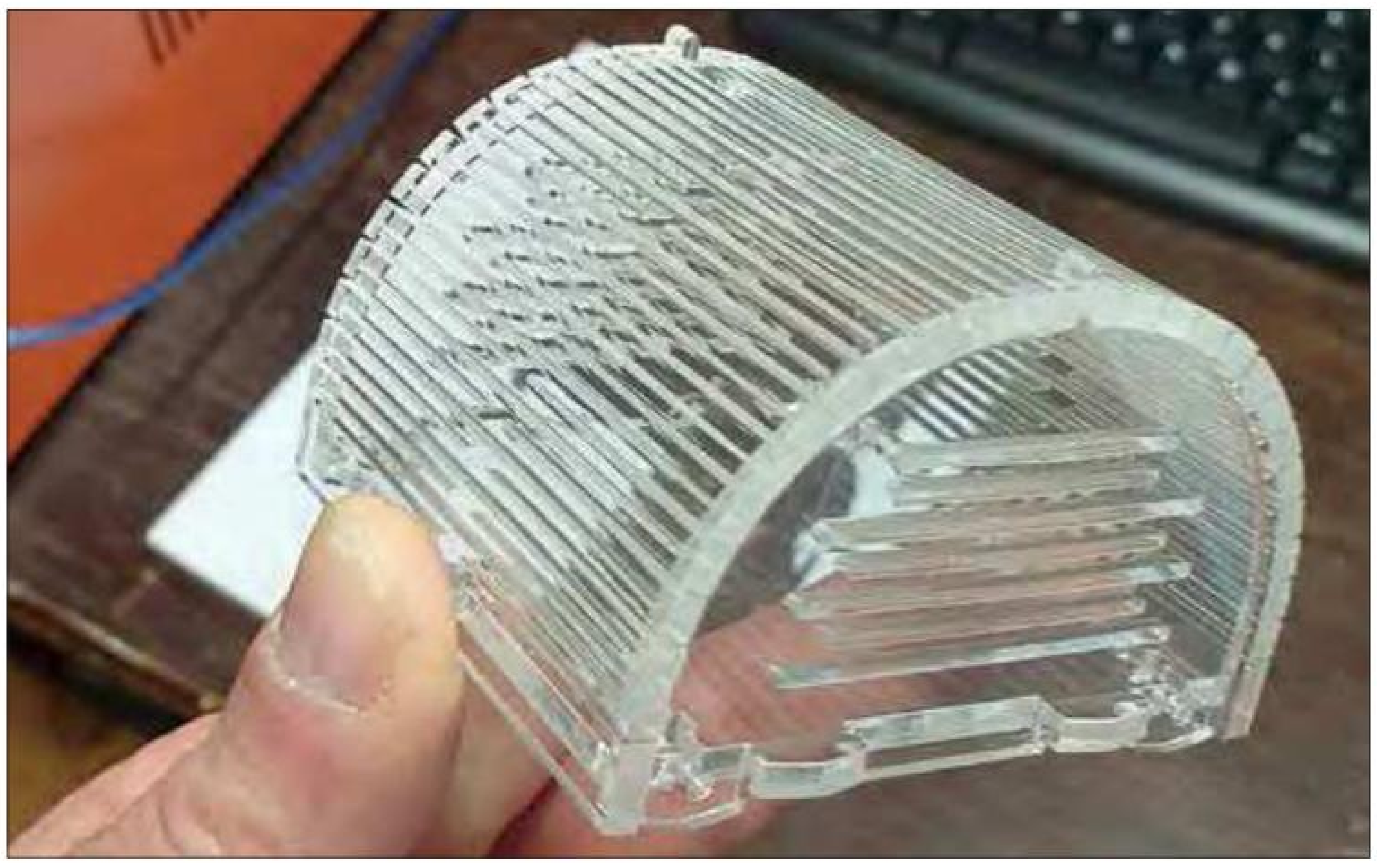


Fig.9. Complex cut-and-fold enclosures are easy to build.

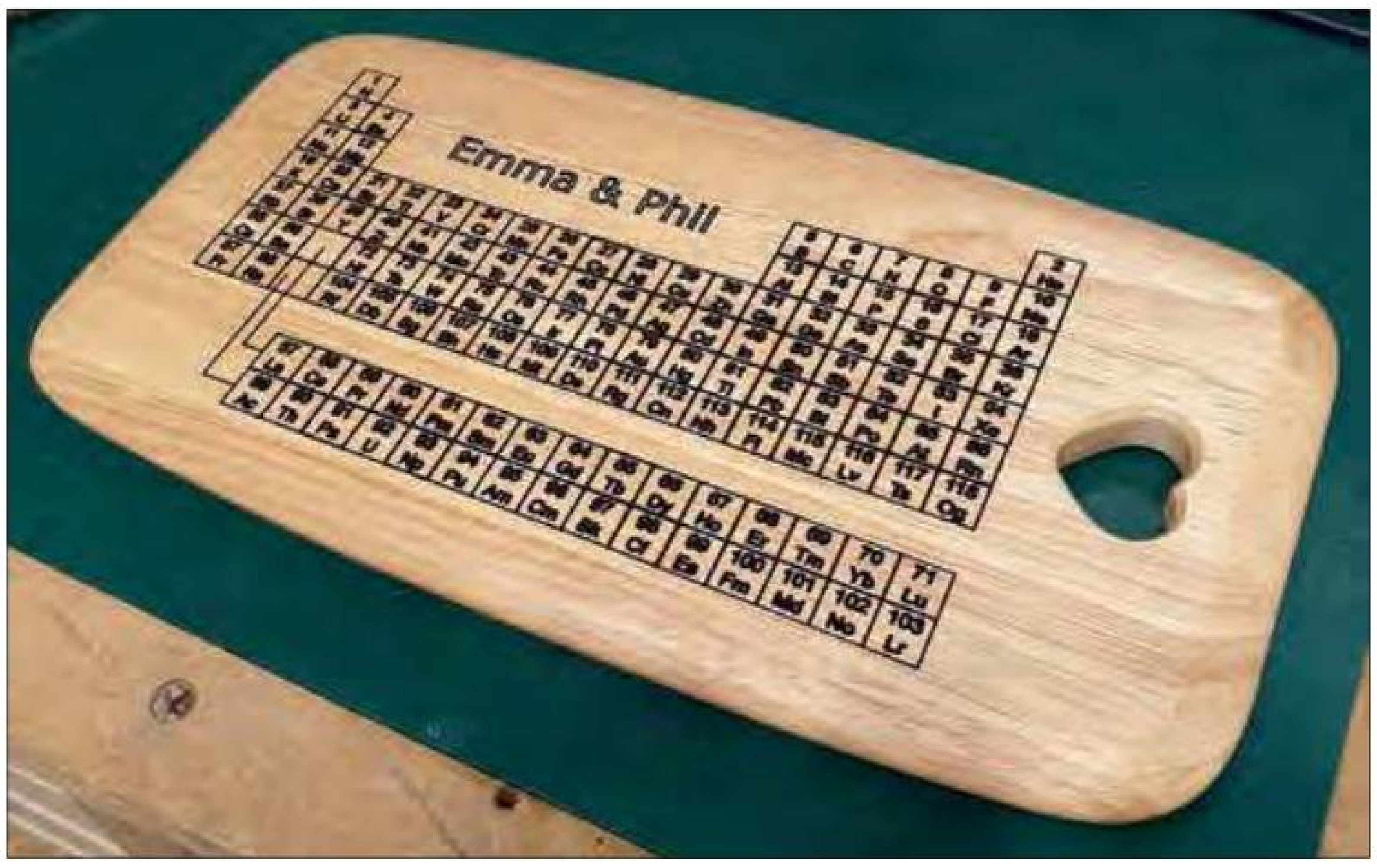


Fig.10. For chemists who cook! A nice example of wood etching.

K40 Whisperer permits three different laser options specified in the image file – cutting, vector engraving and raster engraving. You specify which of these operations to performing by assigning different colours to your image – red for cuts, blue for vector engraving and black for raster engraving. Vector engraving is useful for etching lines and thin text; raster engraving is better for images or thick lines.

In use

So, what can this machine cut? It works with cardboard, 3mm perspex, 3mm plywood and MDF, and leather – we have tested these with success. With multiple passes it is possible to cut thicker materials, but with a reduction in accuracy of the cut. One thing we have found, especially with perspex, is that the cuts are very clean, and the final parts look very professional.

The cut width is approximately 0.2mm, so you can cut very intricate shapes. Unlike 3D printing it's very fast – a 10 × 10 × 5 cm enclosure, consisting of six flat pieces of perspex, took just three minutes to cut.

With so many people using laser cutters in the on-line hobbyist community, there are thousands of free designs available to download and cut or take inspiration from. In Fig.9, for example, you can see an enclosure design we downloaded that uses multiple slits in the top piece to make a curved lid. You are not limited to enclosure and front panel designs; in Fig.10 you can see an etching, that we applied to a bread board for a unique birthday gift. We also created a pair of Christmas earrings – see Fig.11.

In the eight months since it arrived, the K40 has been invaluable – even helping with fast prototyping of an enclosure design for a commercial product: **http://bit.ly/pe-oct20-product**

It's been used to cut custom cardboard packaging designs (including for the above product,) as the power can be controlled accurately enough to etch fold lines in cardboard, not just cut around the edges. Again, software is available for free on the Internet to help create custom packaging designs.

Upgrading

As can been seen from the pictures, the machine is incredibly hackable; indeed, from the start, you have to hack it to make it safe.

Increasing the bed size is the first change, as it is so simple to do and gives the device a massive boost in flexibility to do large jobs. We followed one of a dozen videos on YouTube to create the bed shown in Fig.7 with material sourced at a DIY store using simple tools.

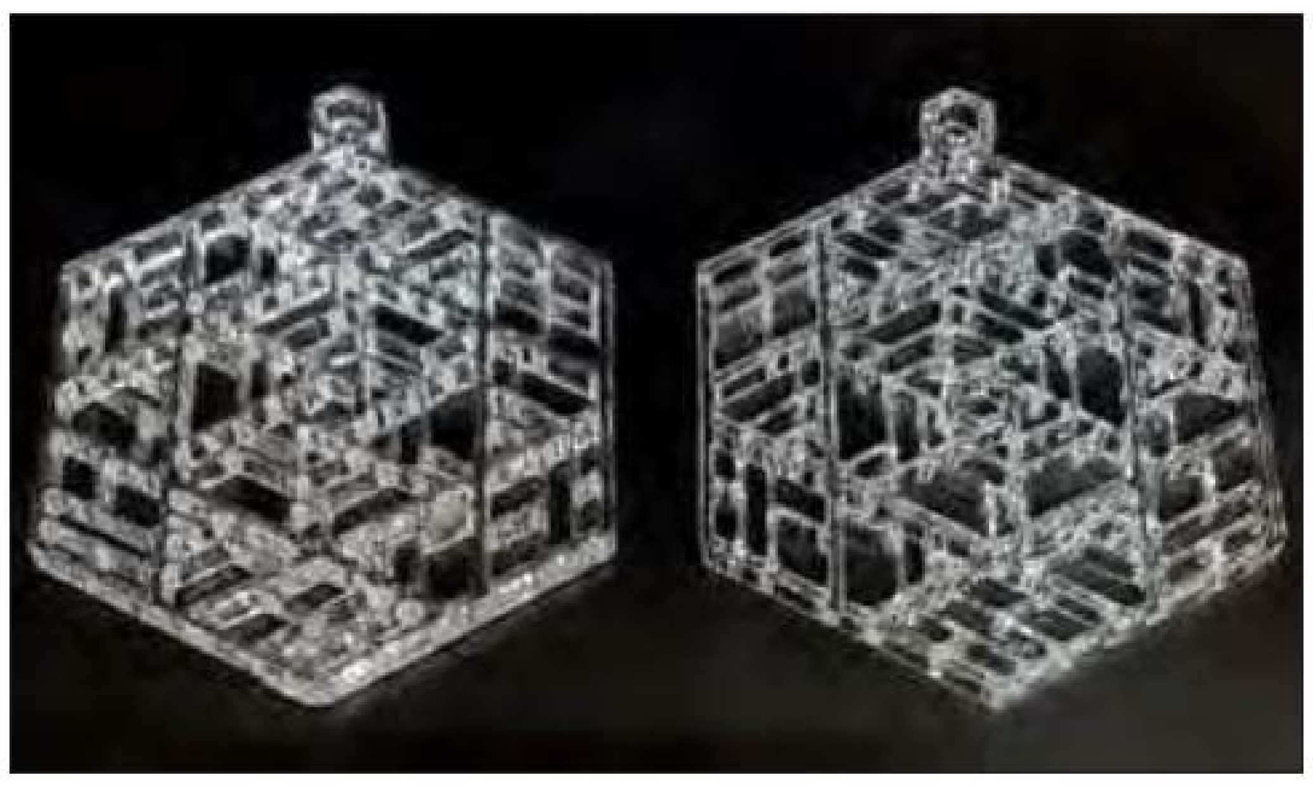


Fig.11. Christmas 'snowflake' earrings.

The next recommended upgrade is air assist; this is an additional air pump that blows air over the target spot on the work piece. This not only reduces burn marks, it makes the cut more accurate and powerful, as the laser beam passes through less pollution before reaching the target. A simple air pump and flexible tubing costs around £20, via eBay.

Everything else can be replaced or upgraded from multiple vendors. It's even possible, with a little more skill, to upgrade the laser tube and power supply to a higher power rating, making it simpler to cut 12mm plastic or wood.

Conclusion – is it time to buy?

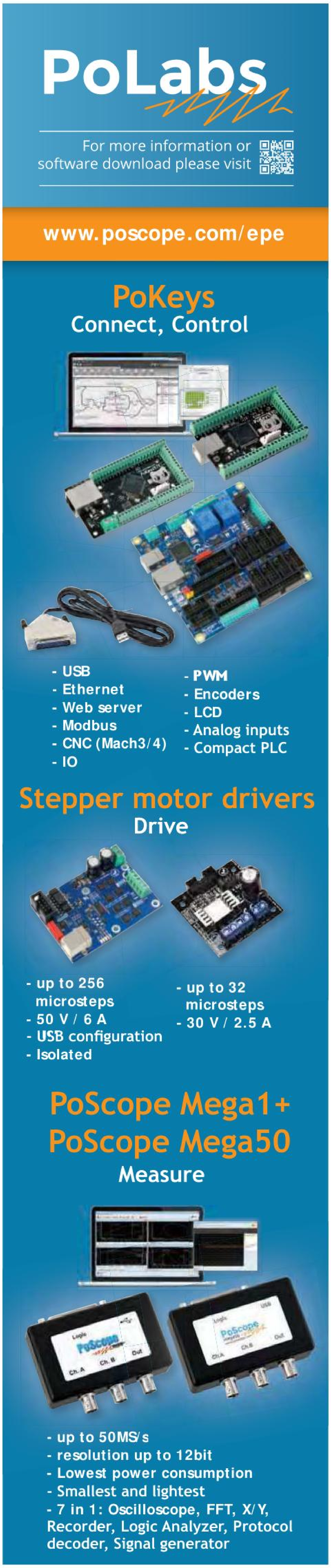
This is a great, affordable tool for the keen hobbyist or maker, but where you live may make its use inappropriate. Without adequate ventilation away from humans and animals, the fumes are too dangerous. We are lucky, our lab is away from other homes and at the end of a garden, with the machine installed beside an exterior wall.

As supplied, the work area is 21cm × 9cm, but easily expandable to 30cm × 21cm, the size of a sheet of A4 paper. This is a perfectly adequate work area for enclosures and instrumentation panels, as well as art and craft cutting and etching.

This is a hugely popular machine around the world, and there are thousands of YouTube videos offering advice on setup, modification and use.

Would we recommend you consider buying one? Yes, but only if you have the space to install it with safe fume venting. Does it replace a 3D printer? No, it complements it.

3D printers have fallen in cost significantly over the years; our original £1000 3D printer was recently replaced with a £140 Ender 3 machine, which performs fantastically. Will we see a similar drop in price for laser cutters? Unlikely in our opinion. First, their use will remain restricted to people with well-ventilated workspaces, limiting broad adoption; second, these machines already benefit from the use of low-cost *x-y* stepper motor control electronics and mechanical solutions developed for the 3D printer market. The main cost in these devices is in the laser tube itself. So, while we will see some reduction in cost over the next decade, don't expect it to be significant.



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Max's Cool Beans

By Max the Magnificent

Flashing LEDs and drooling engineers – Part 8

Beauty and the Beast – but which is which?

I don't know about you, but I can barely keep up with everything that's happening. There's so much fun stuff to do, but never enough time to do it. At the moment, for example, I'm happily experimenting with my 12 × 12 ping-pong ball array, where each ball contains a tricolor LED called a NeoPixel.

In my previous column (*PE*, September 2020), we pretended that the array was lying flat on the floor and that occasional drips of virtual water were randomly falling from the sky. Whenever a virtual drip landed on one of our pixels (ping-pong balls), that pixel lit up bright white for a short period of time.

Next, we decided to represent our drips using random colours selected from a palette comprising three primary colours (red, green and blue), three secondary colours (yellow, cyan and magenta) and six tertiary colours (flush orange, chartreuse, spring green, azure, electric indigo and rose).

The thing is, simply turning our pixels hard on and hard off is not very subtle. In order to add a *soupçon* of sophistication, we decided to fade the colour up, hold it steady, and then fade it back down again, so that's what we are going to do this month.

Keep it simple

In a moment, we're going to create a small suite of functions that will perform the fading effect for us. One trick to writing functions is to make them as simple and as general-purpose as possible. As part of this, it can be advantageous to split parts of our algorithm out into sub-functions that can be reused by other functions in the future.

As usual, it may be advantageous for you to download a copy of this sketch in order to follow along with my meandering musings (just mosey over to the October 2020 page of the *PE* website and download **CB-Oct20-01.txt**).

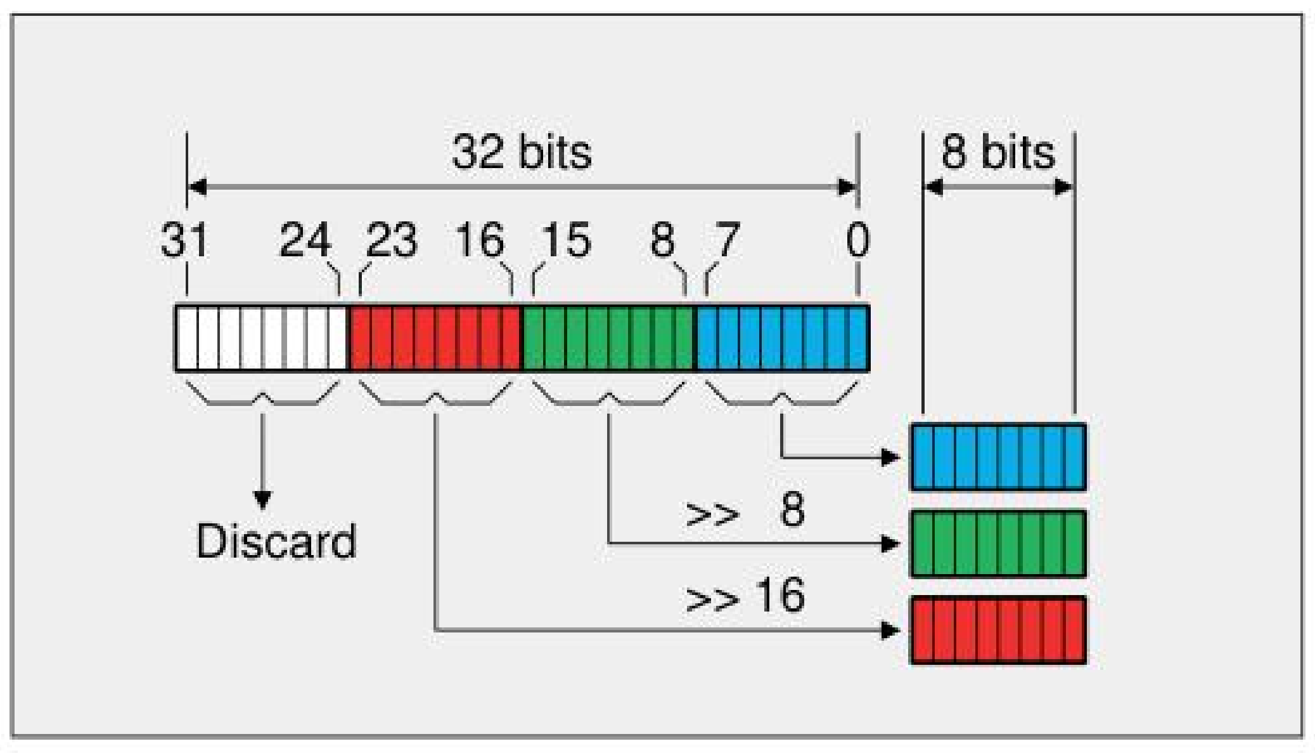

Fig.1. Deconstructing a 32-bit colour value into three 8-bit fields.

The only sizes of fixed-width unsigned integers we have available to us are 8, 16, 32 and 64-bits wide. Our red, green and blue colour channels are each 8-bits wide. As we discussed earlier, we are predominantly representing our colours as single hexadecimal values. For example, we define the `COLOR_WHITE as 0xFFFFFFU` (remember that adding a 'U' or 'u' character on the end of a value directs the compiler to regard it as being unsigned).

Did you spot the fact that `FFFFFF` is only 24 bits wide? Since the smallest unsigned integer we can use to hold this value is 32-bits wide, this means we are leaving the most-significant eight bits unused. Apart from anything else, we should really have specified our `COLOR_WHITE` as `0x00FFFFFFU`. The reason we didn't do so is that we know the compiler will add any required leading zeros when it performs its magic.

Winkle-picking colour channels

Keeping all of this in mind, let's start with three low-level functions called `GetRed()`, `GetGreen()`, and `GetBlue()`. In each case, we are going to pass in a 32-bit colour value and the function will return the appropriate 8-bit colour to us (Fig.1).

Let's start with `GetBlue()`. We pass this a 32-bit value called `tmpColor`. We then use a bitwise `&` (AND) operator to mask out the least-significant eight bits and return the result, which we cast as an 8-bit value (the topic of casting is discussed in this month's *Tips and Tricks* column at the end of this article).

```
uint8_t GetBlue (uint32_t tmpColor)
{
   return (uint8_t) (tmpColor & 0xFFU);
}
```

There are a couple of things to note here. Let's start with the fact that using `(tmpColor & 0xFFU)` to perform the mask operation is the same as saying `(tmpColor & 0x000000FFU)` because, once again, the compiler will insert the leading zeros for us.

More importantly, we don't actually need to perform the mask operation in the first place because the most-significant 24 bits of our 32-bit value will be discarded when we perform the cast. Following on from the previous point, believe it or not, we actually don't even need to use the `(uint8_t)` to cast the result as an 8-bit value, because we already declared the `GetBlue()` function as returning a `uint8_t` value, which means the compiler will automatically perform this cast for us. As a result of all this, we could, if we wished, write our function as follows:

```
uint8_t GetBlue (uint32_t tmpColor)
{
   return tmpColor;
}
```

The advantage of including both the mask and the cast is that it makes our intent clear to anyone who has to understand and maintain this program in the future (remembering that this could well be us). Furthermore, explicitly doing this sort of thing throughout our code can help prevent hard-to-detect bugs from creeping in while we aren't looking.

And, just in case you were wondering (or worrying) about any overhead imposed by these instructions, you may rest assured that the compiler's optimisation knowhow should enable it to recognise any operations that are superfluous to requirements and remove them (if it fails to do so, you can take any additional execution time out of my vacation).

Extracting the blue channel was the low-hanging fruit, because it was already where we needed it to be with regard to its position in our 32-bit colour value. In the case of the green channel and our `GetGreen()` function, we're going to have to shift our 32-bit value eight bits to the right before performing the mask operation.

```
uint8_t GetGreen (uint32_t tmpColor)
{
   return (uint8_t) ( (tmpColor >> 8) & 0xFFU );
}
```

Similarly, in the case of the red channel and our `GetRed()` function, we're going to have to shift our 32-bit value 16 bits to the right before performing the mask operation.

```
uint8_t GetRed (uint32_t tmpColor)
{
   return (uint8_t) ( (tmpColor >> 16) & 0xFFU );
}
```

As an aside, using the & (bitwise AND), | (bitwise OR), and ^ (bitwise XOR) operands is an interesting topic in its own right, especially when it comes to performing masking operations. Sad to relate, we don't have the time to delve into this here, but for those who are excited to learn more, I discuss this in excruciating detail on my *Cool Beans* website – see: **https://bit.ly/3itQGCa**

Regenerating colours

The counterpoint to extracting the red, green and blue channels from a 32-bit colour value is to take a triad, trio, or troika of 8-bit colour channels and use them to construct a 32-bit colour value (Fig.2). We perform this magic using a `BuildColor()` function:

```
uint32_t BuildColor (uint8_t red, uint8_t green,
uint8_t blue)
{
   return ( (((uint32_t) red) << 16) | (((uint32_t)
green) << 8) | ((uint32_t) blue) );
}
```

A little thought shows that we're doing things in reverse to our `GetColour()` functions. First, we cast our 8-bit red channel to be a 32-bit unsigned integer and we shift the result 16 bits to the left. Next, we cast our 8-bit green channel to be a 32-bit unsigned integer and we shift the result eight bits to the left. When it comes to our 8-bit blue channel, all we need do is cast it to be a 32-bit value. Finally, we use | (bitwise OR) operators to merge these three values together to form a single 32-bit colour, which we return to whatever called this function in the first place.

Watch me fade!

For reasons that will become apparent in a future column, we're going to implement a 'master clock' with a duration of 10ms (milliseconds). This is defined by `TICK` in our sketch. Purely for the sake of discussion, let's suppose we decide to take 100ms to perform our fade. Using our 10ms master clock, this means we're going to require 100/10 = 10 steps

When it comes to actually performing the fade, the next function we require is one that can calculate a new colour value that's formed as a specified proportion of two different colours. Also, just for giggles and grins, we will need to perform this calculation on each of the red, green and blue channels independently.

If you look at our sketch, you'll see a function called `CrossFadeColor()`. When we call this function, we pass in four arguments, the:

- 32-bit `startColor`
- 32-bit `endColor`
- Total number of steps in this fade: `numSteps`
- Step we're currently on `currentStep`.

Let's look at the statement we use to process the red channel as follows (observe that this statement calls our `GetRed()` function two times to extract the red channels from `startColor` and `endColor`):

```
tmpRed = ((GetRed(startColor) * (numSteps –
currentStep)) + (GetRed(endColor) *
currentStep)) / numSteps;
```

Now, this can be a little tricky to wrap one's brain around. The best way for you to comprehend the nuances of all this will be to draw a table that contains columns for each of the possible `currentStep` values, which would be 0 to `numStep` (ie, 0 to 10). Your table should also contain four rows; one each for the:

- Results of the `startColor` calculations
- Results of the `endColor` calculations
- Results of adding these two values
- Ultimate results following the final division operations.

I suggest you begin by assuming that the red component of `startColor` has a value of 0 (fully off), while the red component of `endColor` has a value of 255 (fully on). Next, create a second table using a `startColor` value of 255 and an `endColor` value of 0. Finally, do the whole thing again with both values set to 255, which could easily happen if the main colours were magenta (red and blue) and yellow (red and green).

Do you recall earlier on when I said, 'One trick to writing functions is to make them as simple as possible'? I don't know about you, but I think the calculation we just introduced is rather cunning. If you look at our `CrossFadeColor()` function, you'll see that it contains only four lines of code that are doing the 'heavy lifting.' Three

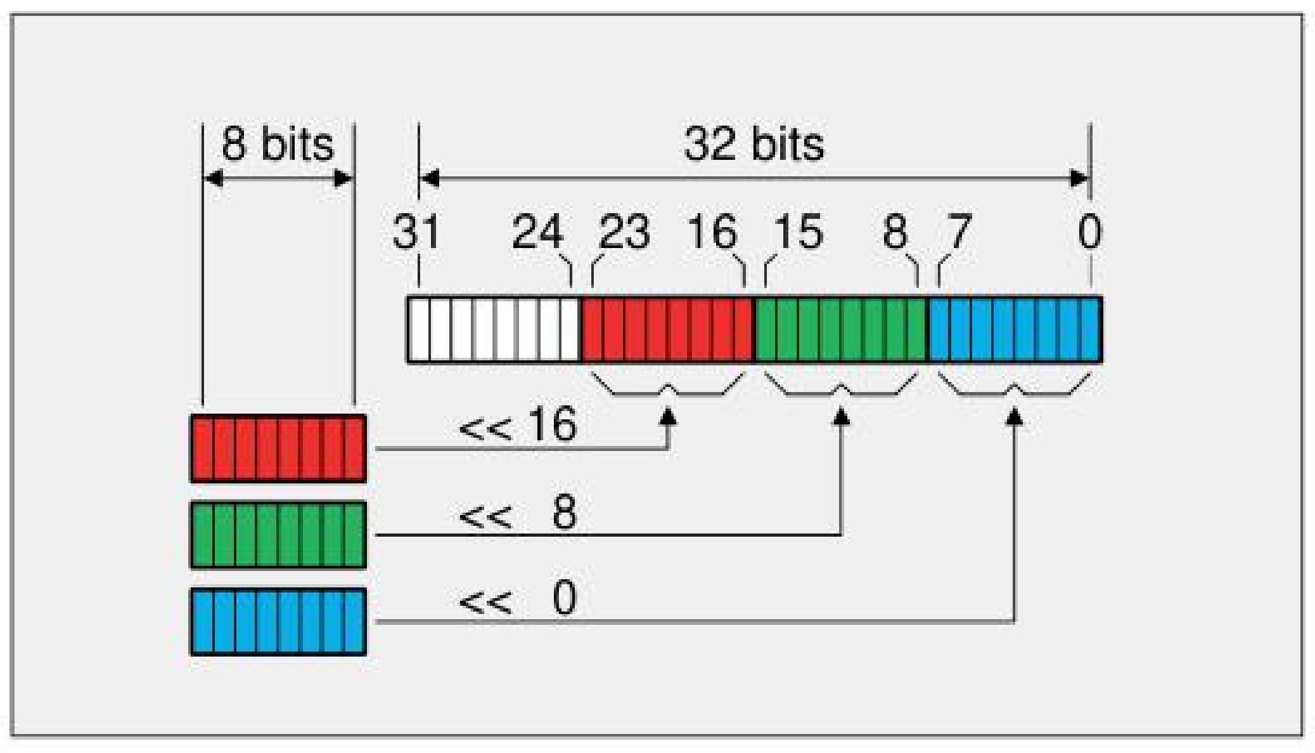

Fig.2. Using three 8-bit fields to construct a 32-bit value.

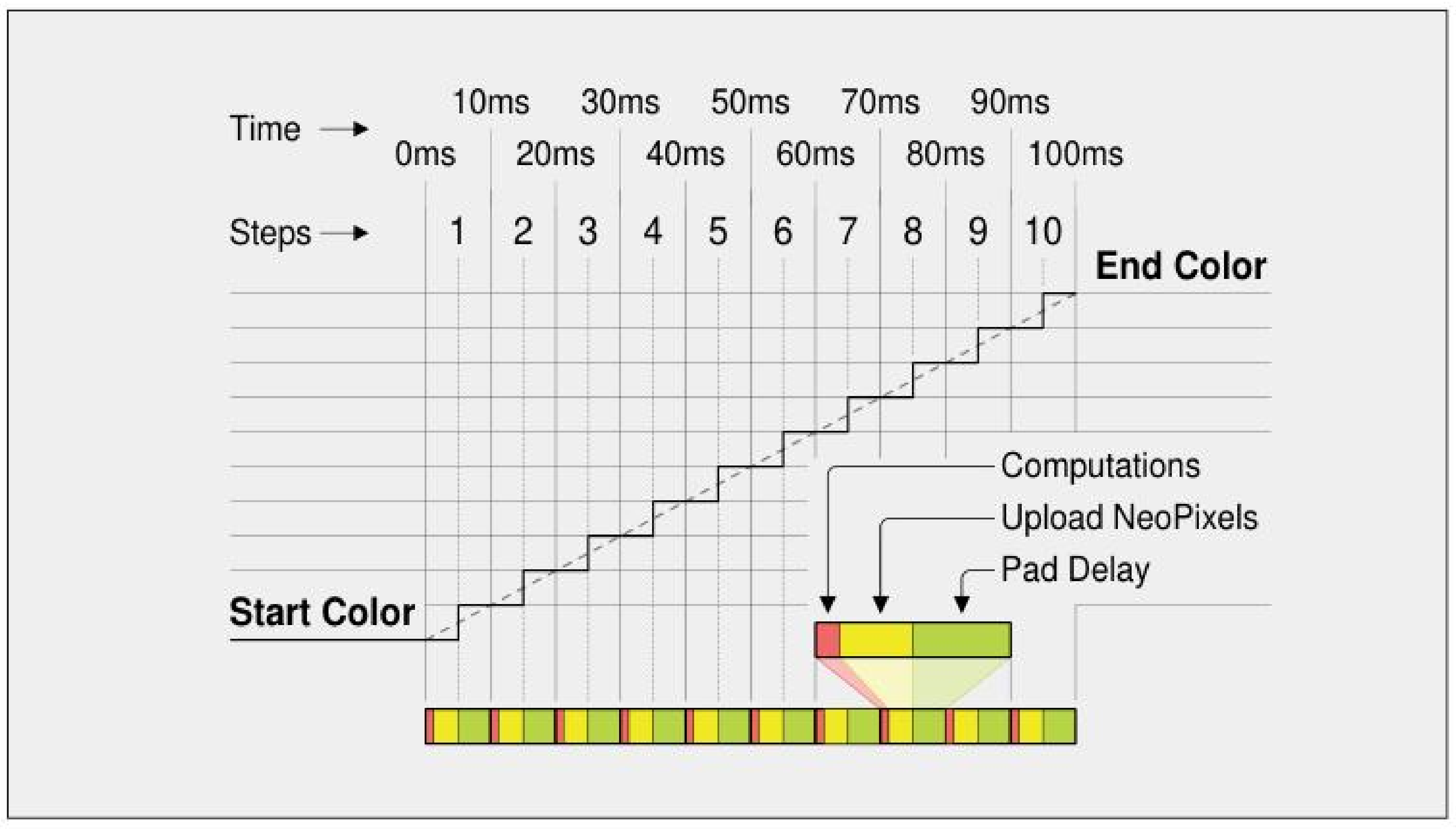

Fig.3. Graphical representation of 100ms fade with 10ms master clock.

of these lines calculate the new red, green, and blue channel values, while the fourth calls our `BuildColor()` function to merge everything back together again.

All that is required now is one more function we'll call `CrossFade()`, which we'll use to orchestrate our fade effect. If you look at the sketch, you'll see that the first thing we do in this function is to calculate the number of fade steps (`numFadeSteps`) based on the required fade duration and the value of our master clock tick. The next thing we do is to perform a loop that calls our `CrossFadeColor()` function as follows:

```
for (int iStep = 1; iStep <= numFadeSteps; iStep++)
{
   fadeColor = CrossFadeColor(startColor,
endColor, numFadeSteps, iStep);
   // More stuff here
}
```

Before we proceed, let's pause to ponder the fact that there are four main options for the range of values we could assign to `iStep` to control our loop:

- `0 to < numFadeSteps`
- `0 to <= numFadeSteps`
- `1 to < numFadeSteps`
- `1 to <= numFadeSteps`.

So, why did we choose the latter option? Well, when it comes to this sort of loop construct, the initial value and terminating condition will depend on our algorithm and what we're trying to do with it. There's a common issue in computer programming that's called the 'off-by-one error' or 'off-by-one bug.' This is also commonly known as a 'fence post error' based on the so-called 'fence post problem.' For example, assuming we have fence posts spaced 10 feet apart, how many posts will be required to support a 100-foot length of fence? The answer is 11 – there is always one more fence post than there are fence spans – because we need a post at the beginning of the fence. Another way to look at this is that a *single* 10-foot length of fence will require *two* posts – one at either end. A similar condition can occur in our programs when an iterative loop iterates one time too few or one time too many.

I'm sure professional computer programmers have no problem with this sort of thing, but it typically makes my head hurt. On the bright side, I'm a visually oriented person, so I usually find it helps me to draw things out. Let's create a graphical representation of our 100ms fade with a 10ms master clock (Fig.3).

Before we commence our fade, we can assume that we already have 100% of the original (start) colour and 0% of the new (end) colour. By the time we finish the fade, we wish to have 0% of the original colour and 100% of the new colour. The 'per cent' values of the intermediate colours will depend on the current step when compared to the total number of steps.

As defined by the NeoPixel library, the clock used to upload our NeoPixel string is running at 800kHz. Each NeoPixel requires 24 bits of data, which takes (1/800,000) * 24 = 30µs. Since we have 145 NeoPixels (including our 'sacrificial pixel'), this equates to a total delay of 30 * 145 = 4,350µs = 4.35ms. If we 'guestimate' that all of our computations require 0.65ms (we will return to consider this in more detail in a future column), then this means we need to add a padding delay of 10 – (0.65 + 4.35) = 5ms to each of our steps to build things up to our 10ms master clock tick (you will see this defined as `INTER_TICK_PAD_DELAY` in our sketch).

Once again, do you recall earlier on when I said, 'One trick to writing functions is to make them as general-purpose as possible'? Well, the beauty of our `CrossFade()` and `CrossFadeColor()` functions is that it doesn't matter whether we are 'fading up' (from black to a colour) or 'fading down' (from a colour to black), because black is just another colour. In turn, this means that we can use these functions to fade from any colour to any other colour.

For your delectation and delight, I've created a video showing our colour-fading sketch in action (**https://bit.ly/2EXQ20k**). In my next column, we will delve more deeply into various multi-colour combinations we might decide to employ. Until that frabjous day, please Callooh! Callay! – in other words have a good one!

Cool bean Max Maxfield (Hawaiian shirt, on the right) is emperor of all he surveys at **CliveMaxfield.com** – the go-to site for the latest and greatest in technological geekdom.

Comments or questions? Email Max at: **max@CliveMaxfield.com**

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Max's Cool Beans cunning coding tips and tricks

On the one hand, I keep on saying I'm a hardware designer by trade and my software skills are rudimentary at best. On the other hand, I constantly amaze myself with all the tidbits of trivia and nuggets of knowledge that have managed to take root in my poor old noggin.

Want an argument?

As one example of the above, someone just emailed me to ask why I sometimes use the term 'parameter' and at other times say 'argument'. Well, consider the following declaration of a meaningless function:

```
int SillyFunction (int a, int b)
{
   int y = a + b;
   return y;
}
```

When we declare a function, part of this declaration is a list of *parameters* associated with the function. In this case, our function has two parameters: `a` and `b`. Of course, it's also possible for a function to be declared with an empty parameter list.

Now consider when we call our function from somewhere else in the program; for example:

```
int sillyResult = SillyFunction(40, 2);
```

When we call a function, we pass in a list of *arguments*. In this case, we pass two arguments, `40` and `2`, into `SillyFunction()`. Of course, if a function is declared with an empty parameter list, then when we call it we will pass in an empty argument list.

In the real world, a lot of people use the terms 'parameters' and 'arguments' interchangeably. So long as the person you are talking to understands the message you are trying to convey, then there's 'no harm, no foul,' as they say.

The problem comes when you are talking with professional programmers who will take great delight in painstakingly instructing you in the error of your ways. Personally, I don't care to give the little rascals the satisfaction.

Curiouser and curiouser

Towards the end of *Chapter 1* in *Alice's Adventures in Wonderland*, Alice foolishly guzzles the contents of a small bottle marked 'Drink Me' and shrinks to only ten inches in height. A little later, the little scamp cannot restrain herself from munching down on a small cake with 'Eat Me' marked in currants. At the start of *Chapter 2*, Alice cries 'Curiouser and curiouser!' as she rapidly grows to giant size. If this ever happens to me (again), I'm sure I will say much the same (or words to that effect).

The reason I'm waffling on about this here is that something similar can happen with integers in C/C++ (actually, it can happen with all data types, but we will focus on integers for the purpose of these discussions). This is known as 'type conversion', of which there are two flavors: *implicit* and *explicit*. In the case of implicit-type conversion, the compiler accomplishes this automatically without our having to instruct it to do so. Alternatively, we can use explicit type conversion to 'cast' (transform) a value from one data type to another.

If you take a look at last month's *Tips and Tricks* column (*PE*, September 2020), you'll see that we have short, regular and long versions of signed and unsigned integers (we also have fixed-width data types like `uint8_t` and `uint32_t`). Whenever we try to perform a binary operation on operands of different types, the compiler will automatically convert one of the operands into the same type as the other. If the operand goes from a smaller domain (eg, short) to a larger domain (eg, long), this is called *promotion*. By comparison, if the operand goes from a larger domain to a smaller domain, this is known as *demotion*.

Promotion typically isn't a problem because the range of values associated with the smaller integer forms a subset of the larger domain to which it is being promoted. In the case of demotion, however, problems may occur if the value being demoted is too large to fit into the target type, in which case the result will be truncated (this may, of course, be just what we want to occur).

Different languages deal with all of this in different ways. In the case of C/C++, we can perform explicit-type conversion using a cast operator, which involves the name of the new data type surrounded by parentheses. For example, suppose we declare two unsigned integer variables: one 8-bits wide and the other 32-bits wide, as follows:

```
uint8_t  myInt8;
uint32_t myInt32;
```

Now suppose that, somewhere in our program, we wish to copy the contents of the 32-bit value into the 8-bit value. In order to achieve this, we would cast the 32-bit value into its 8-bit counterpart as follows:

```
myInt8 = (uint8_t) myInt32;
```

One of the purposes of the cast operator is to inform the compiler that we know what we're doing. In this case, the most-significant 24 bits of the 32-bit `myInt32` will be discarded, and only its least-significant eight bits will be copied into `myInt8`. As we've seen, the cast operator has only one operand, which is to the right of the operator. As always, problems lurk for the unwary. Suppose we want to shift the value in `myInt32` eight bits to the right and then copy the least-significant eight bits of the result into `myInt8`. Consider the following statement:

```
myInt8 = (uint8_t) myInt32 >> 8;
```

What do you expect this to do? Many beginners would expect this to first perform the shift operation followed by the cast. In reality, the cast operator has a higher precedence than the bitwise shift operator, so the operation that will actually be performed can be represented as follows:

```
myInt8 = ((uint8_t) myInt32) >> 8;
```

That is, the 32-bit value will first be cast into an 8-bit value, which will subsequently be shifted eight bits to the right. The result will be to leave `myInt8` containing `0` (`00000000` in binary, `0x00` in hexadecimal), which is not what we were hoping for. In order to achieve the desired result, we would actually have to write our statement as follows:

```
myInt8 = (uint8_t) (myInt32 >> 8);
```

Adding these parentheses forces the shift to be implemented first, after which the cast is performed on the result. You may not be surprised to learn that the reason we are talking about this here is that in my *Cool Beans* columns we are poised to start casting like champions (I've just dispatched the butler to fetch my casting trousers).

Make it with Micromite

Phil Boyce – hands on with the mighty PIC-powered, BASIC microcontroller

Part 21: Exploring a GPS module with UART communication

This month we will explain how to connect the Micromite to external hardware that uses an interface commonly known as 'UART'. 'UART' stands for 'Universal Asynchronous Receiver-Transmitter', which in essence relates to data that is transmitted (and received) between two devices using a specific serial protocol. There are many readily available hardware modules that use this protocol, and once you have an understanding of the basic concepts it will be easy to implement UART-based hardware in future Micromite projects.

Throughout this series we have tried to keep things simple by demonstrating new topics with just a few lines of code, which in turn interface with minimal hardware. Here, we will again adopt this approach by first covering the concept of UART communication, and then introducing the relevant MMBASIC commands. With these two elements understood, we can then explore how to communicate with a simple UART module.

To make this fun, our aim will be to listen to, and display, the serial data 'sentences' that a low-cost Global Positioning System (GPS) receiver module continuously outputs. We will show you how to use this data in various ways, including plotting your position on a map, and how to build a simple, yet highly accurate clock based upon the precise time information that is contained within GPS data.

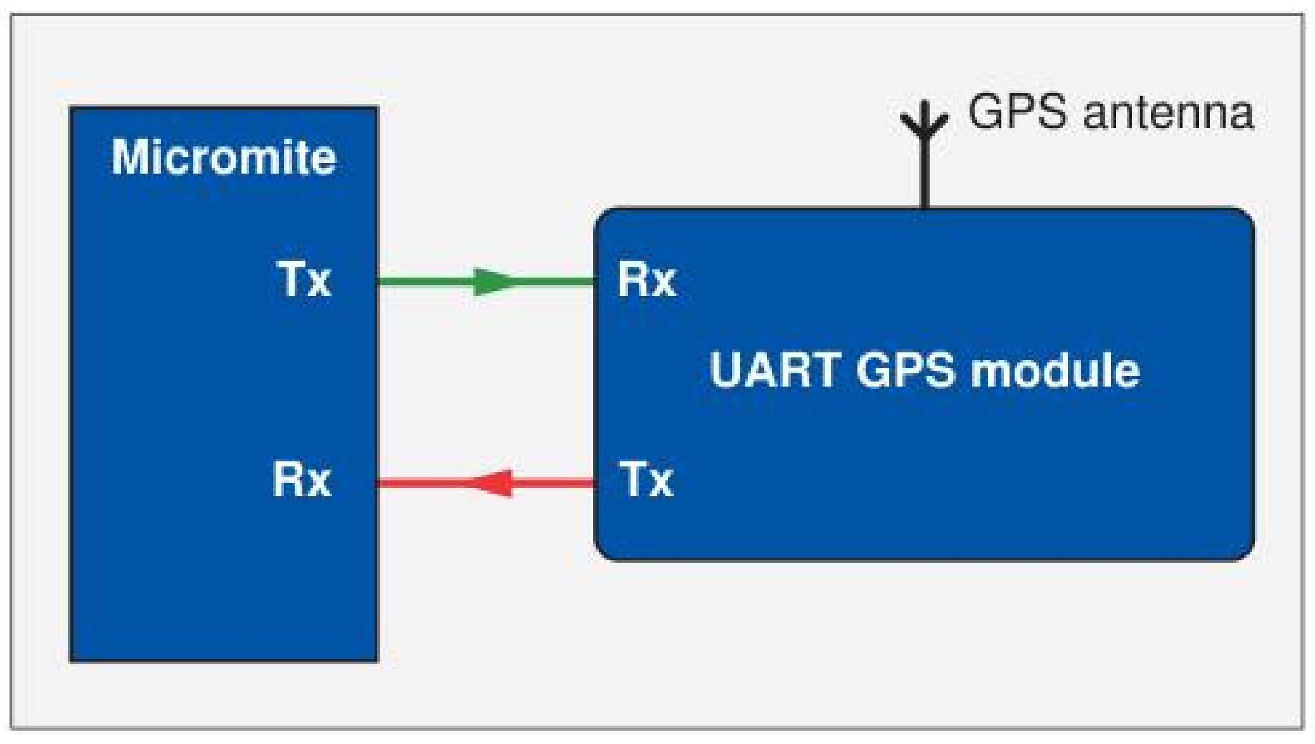


Fig.1. Concept for connecting any UART module (eg GPS) to the Micromite.

UART concepts

Back in *Part 11* (*PE*, December 2019) we discussed the theory of serial data communication. That included a brief description of how asynchronous communication works between two devices. If you have access to *Part 11*, then it is well worth rereading. However, if you don't have that issue, here's a brief summary of the concepts.

For two devices to communicate serially with each other, the transmit pin (Tx) of the first device needs to be connected to the receive pin (Rx) of the second device (and *vice versa*: the Tx on the second device connected to Rx on the first device). If we now consider the first device as being a Micromite, and the second device as a UART GPS module, then this setup represents how the hardware we'll be using this month is required to be connected – see Fig.1.

An important point to clarify here is that the two 'signal-lines' shown in Fig.1 are each their own dedicated communication link and they work independently of each other. Each communication link has a transmitter at one end, and a receiver at the other. Tx is the 'source' of the data, and the Rx is the 'destination' of the data; hence the green link is used when the Micromite wants to 'talk' to the GPS module (which in effect is 'listening' on its Rx pin); and the red link is used when the GPS module wants to 'talk' to the Micromite.

In our application here, we only need the Micromite to listen to data being outputted from the GPS module. Therefore, we only need the red link shown in Fig.1.Thus, the GPS module's Tx pin will effectively output serial data which is received at the Micromite's Rx 'listening' pin.

That covers the physical connections, now let's look at various elements of the UART protocol itself – ie, how the communication takes place.

UART protocol: Baud-rate

Asynchronous communication requires that both devices (at either end of a communication link) operate at exactly the same speed as each other. This is because the transmitter needs to set the communication link to either a logic high or a logic low for a defined amount of time, during which the receiver then needs to sample the logic level present on the communication link. Think of this as a *Tx-set/Rx-sample* process representing the communication of a single bit of information. This process must be repeated at regular (ie, agreed) time intervals in order for each bit of transmitted data to be received correctly by the receiver and formed back into a meaningful piece of data (comprising a certain number of bits). If the two devices are not operating at the same speed, then very quickly the Tx-set (and/or the Rx-sample) will occur at the wrong time and hence corrupt data will be received.

Consider the transmission of a single byte of data, which effectively means the transmission of 8-bits. In order for the byte to be successfully communicated asynchronously from the transmitting device to the receiving device, the Tx-set/Rx-sample process needs to be repeated eight times at precise time intervals. This time interval is defined by what is known as the UART 'baud rate'. We discussed this in *Parts 3 and 4* (*PE*, April and May 2019) when we talked about how to set up your Terminal application (we used TeraTerm) to successfully communicate with your

Micromite code

The code in this article is available for download from the *PE* website.

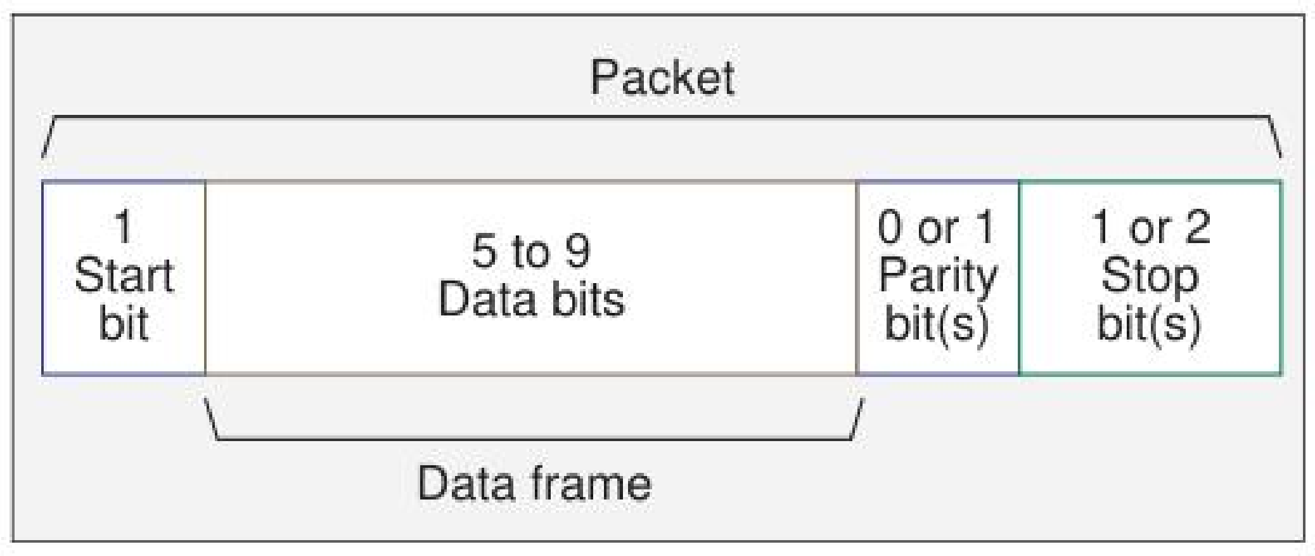


Fig.2. UART communication involves sending packets. The required data is 'wrapped' with a Start bit, a Parity bit, and 1 or 2 Stop bits, as shown here.

MKC. Note that the baud rate needs to be set 'manually' at both ends; and if set to different values then communication will simply not work correctly.

UART protocol: Start bit

There is one other important detail for successful UART communication to take place, and that is how does the Tx device and Rx device know when to start the Tx-set/Rx-sample process. For example, if you were to power up only the transmitter circuit and let it begin the Tx-set part of the process, then there would be nothing 'listening' on the other end of the communication link to receive (ie, sample) the data. If you now power up the receiver circuit, then the Rx-sample process is potentially occurring part way through a message that is currently being transmitted. Hence, in this scenario, the received data would be incorrect and/or incomplete.

To solve this issue, the UART protocol simply 'wraps' the data into 'packets', with the first bit of a UART packet being a 'Start Bit' indicator (see Fig.2). It is this Start Bit that synchronises the receiver with the transmitter so that the data is received correctly.

In summary, a UART transmitter (Tx) which is at one end of a communication link, is responsible for sending the data to the receiver (Rx) which is at the other end of the communication link. The sending/receiving of data is implemented with the Tx-set/Rx-sample process on the communication link. The Start Bit in the 'UART packet' synchronises the Rx with the Tx; and the transmitter and receiver need to be set to operate at the same speed (baud rate).

The UART packet

Referring to Fig.2, we can see that there are four main sections to the UART packet. To be able to use UART communication in its simplest form, we need a basic understanding of all four sections.

1. The Start bit synchronises the Tx-set/Rx-sample process. Think of it as Tx telling Rx to start listening.
2. The Data frame contains the *actual* data being transferred serially. It can comprise 5, 6, 7, 8, or 9 bits (8 is the most common since it makes a byte of data).
3. The Parity bit provides a partial method to determine if the data has been received accurately. Three terms associated with this include: no parity, odd parity, and even parity. If parity is used, then the total number of '1' bits in the data frame are counted by the Tx, and then the parity bit is either set, or left unset by the Tx so that there is either an odd number, or even number of '1' bits in total (in the data frame *and* the parity bit). The Rx then uses this parity bit to check that it also receives an odd or even number of '1' bits. Note that parity checking can only identify *some* data-error scenarios, but not all.
4. Stop bits (either one or two) are sent by the transmitter to signify to the receiver that the transmission is complete.

The point of highlighting the above is that there are several parameters that need to be defined when using UART communications. These are then set as required in order to configure both devices at either end of the communication:

- Baud rate: defines the speed of data transmission – here we will be using 9600 baud
- Data bits: between 5 and 9 bits can be used – here we will be using 8 data bits
- Parity: N(o), O(dd), E(ven) – here we will be using No parity
- Stop bits: 1 or 2 – we will be using 1.

The above is often seen written as something similar to: 8-N-1 (9600). Without knowing (and setting) these basic parameters, UART data communication will not work correctly (if at all).

There are several more additional elements to UART communication and a quick search on Google will allow you to read more if you're interested. For example, the signals can be logically inverted; however, the point in this article is to keep things to the bare minimum so we can understand how to use MMBASIC commands to allow communication with a UART module.

Micromite UART pins

The 28-pin Micromite (as used in the *MKC*) has a total of three hardware UART ports. These are referred to as: 'COM1', 'COM2', and 'COM3'. Each has a Tx pin and an Rx pin. COM3 is specifically used as the console connection – ie, to connect the *MKC* to the Terminal App (or more specially, COM3 is connected to the *Development Module*). However, COM1 and COM2 are both available for you to connect to UART hardware. Refer to Fig.3 to see the relevant Micromite pin numbers. In our application, we will be connecting the GPS module to COM1 Rx.

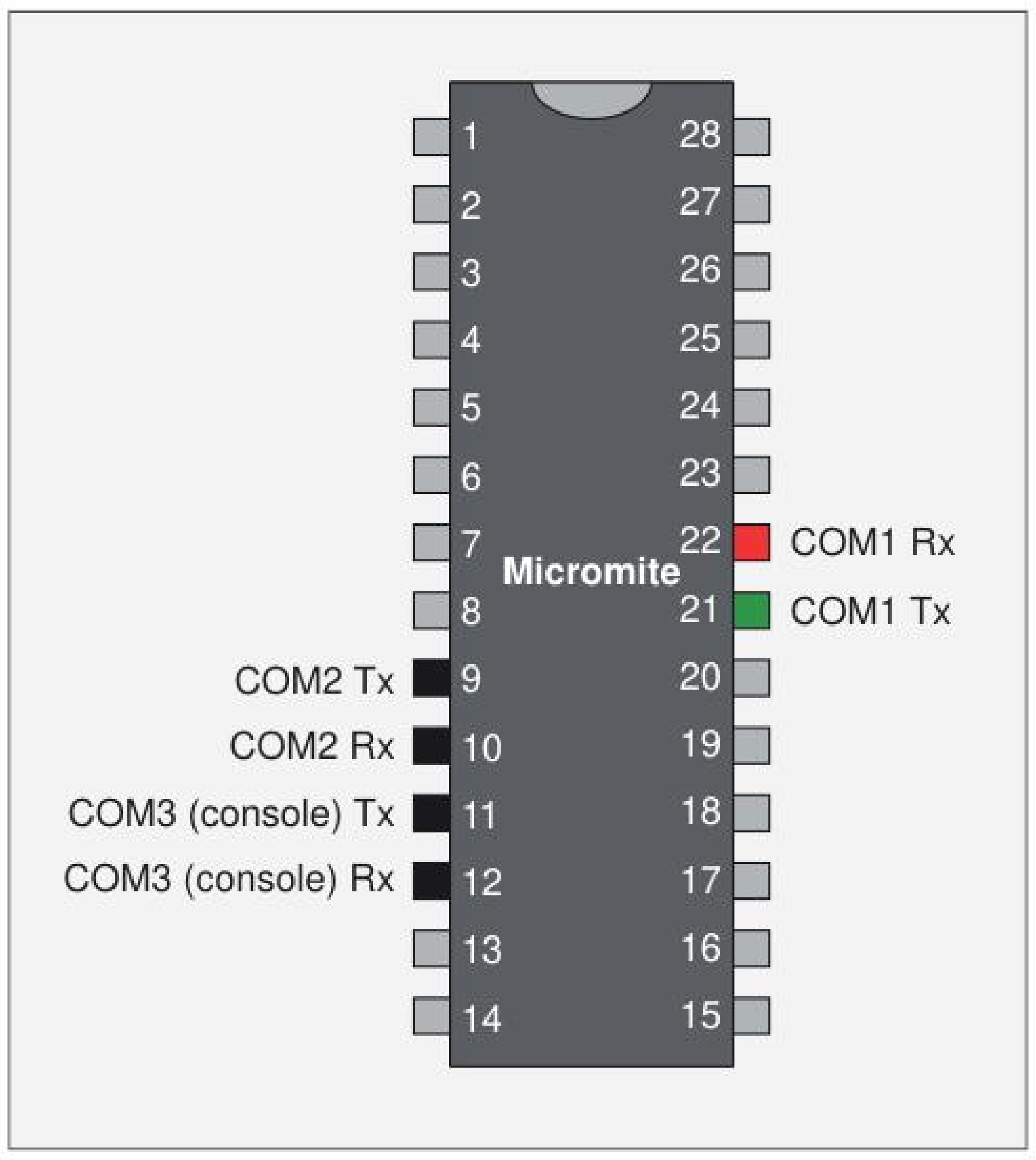


Fig.3. The Micromite used in the *MKC* has three UART (COM) Ports. COM3 is reserved for the Console connection.

Now that we understand the concept of UART communication, along with knowing what certain UART terminology refers to, we are almost ready to connect a GPS UART module to our *MKC*. However, we first need to select a suitable GPS module from the many types available.

GPS module choice

A GPS receiver module comprises two main parts: the antenna (that receives the GPS signal), and the GPS receiver itself. The GPS receiver contains all the necessary electronics, and these will typically be mounted on a small PCB. The antenna may be part of the PCB, or instead the PCB will allow for an external antenna to be connected.

The PCB will also have either a row of header pins, or allow for a small connector/lead to be plugged in. This is the physical connection to the GPS module, allowing it to be powered and use the communication link. Note that the module may communicate over various protocols, including UART, I^{2}C and SPI.

A quick online search for 'UART GPS module' will highlight that there are many different types of GPS modules available. They can start at a cost as low as £3 (but we would advise against these) and a high-end GPS module can cost £50 or more. You may even have a suitable GPS module already in your collection of 'useful bits for a rainy day'.

We recommend a GPS module that comes with an external antenna, and ideally one that is based on a recognised GPS receiver module such as the popular U-Blox 6M, or SIM28. The one we have used in this article is the Grove GPS module (see Fig.4). It is a low-cost, high-performance module that is readily available for around £20 from many recognised suppliers such as RS and Farnell (as well as from many smaller online suppliers). If you do an online search you will see potential

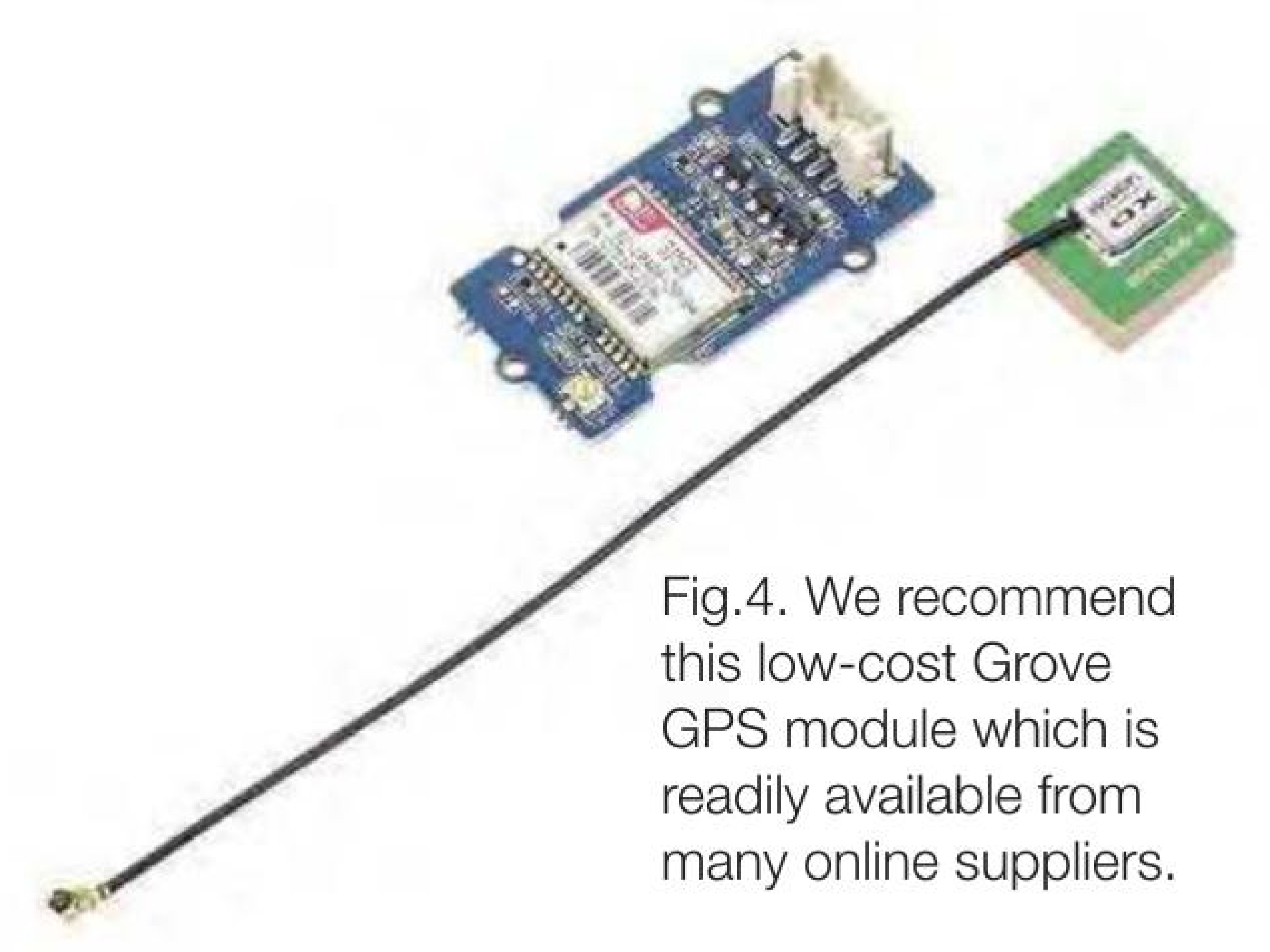


Fig.4. We recommend this low-cost Grove GPS module which is readily available from many online suppliers.

suppliers that can send you one at a very reasonable cost.

The issue with most of the cheaper units typically available from popular auction websites is that they often use a sub-standard antenna and poor-quality antenna lead. This results in a GPS receiver that can't actually receive the GPS signal. Either that, or you have to position the module outdoors just to be able to receive the GPS signal. The Grove GPS module on the other hand has been tried and tested indoors and works very well. However, do take care when handling the uFL antenna connector/lead – it is rather small and delicate! (See Fig.5, left – by the way, we have no affiliation to the manufacturer of this particular module.)

Another point to consider is using a uFL-to-SMA adaptor in order to use an antenna that has a much longer lead – see Fig.5, right. Antennae with longer leads typically have an SMA connector and hence the need for the adaptor. Using such a setup allows you to have the GPS antenna near the window, which improves reception of the GPS signal. It also avoids having to move your computer!

GPS circuit diagram

The circuit diagram for connecting a GPS UART module to the Micromite is shown in Fig.6. As usual, it is just a matter of connecting the correct pins together between the GPS module and the *MKC*. This could hardly be any simpler since just three connections are required: two for supplying power to the module, and one for the communications link.

Most GPS modules operate from a 5V supply, but do be careful to check your module's spec, just in case it is a 3.3V module. If so, then simply connect to the Micromite's 3.3V output.

Due to the simplicity of the circuit, we are not using stripboard; instead we will just use three male-male jumper wires, each plugged between the *MKC* and the supplied GPS connector lead.

If you have been following this series and have built the Bluetooth (BT) module, then we would advise using it. This allows the *MKC* (and BT module) to be connected to the GPS module (see Fig.7) and then everything positioned near to a window with a 'view to the sky', which allows better GPS reception.

However, if you have not built the BT module then simply plug the *DM* into your *MKC*, and connect the GPS as shown in Fig.6.

MMBASIC UART commands

Appendix A of the *Micromite User Manual* is dedicated to UART communication and is worth reading (see this month's download). However, to keep things simple we will just highlight the single command that correctly configures the specific UART port to which we have connected the GPS module to. The command line is:

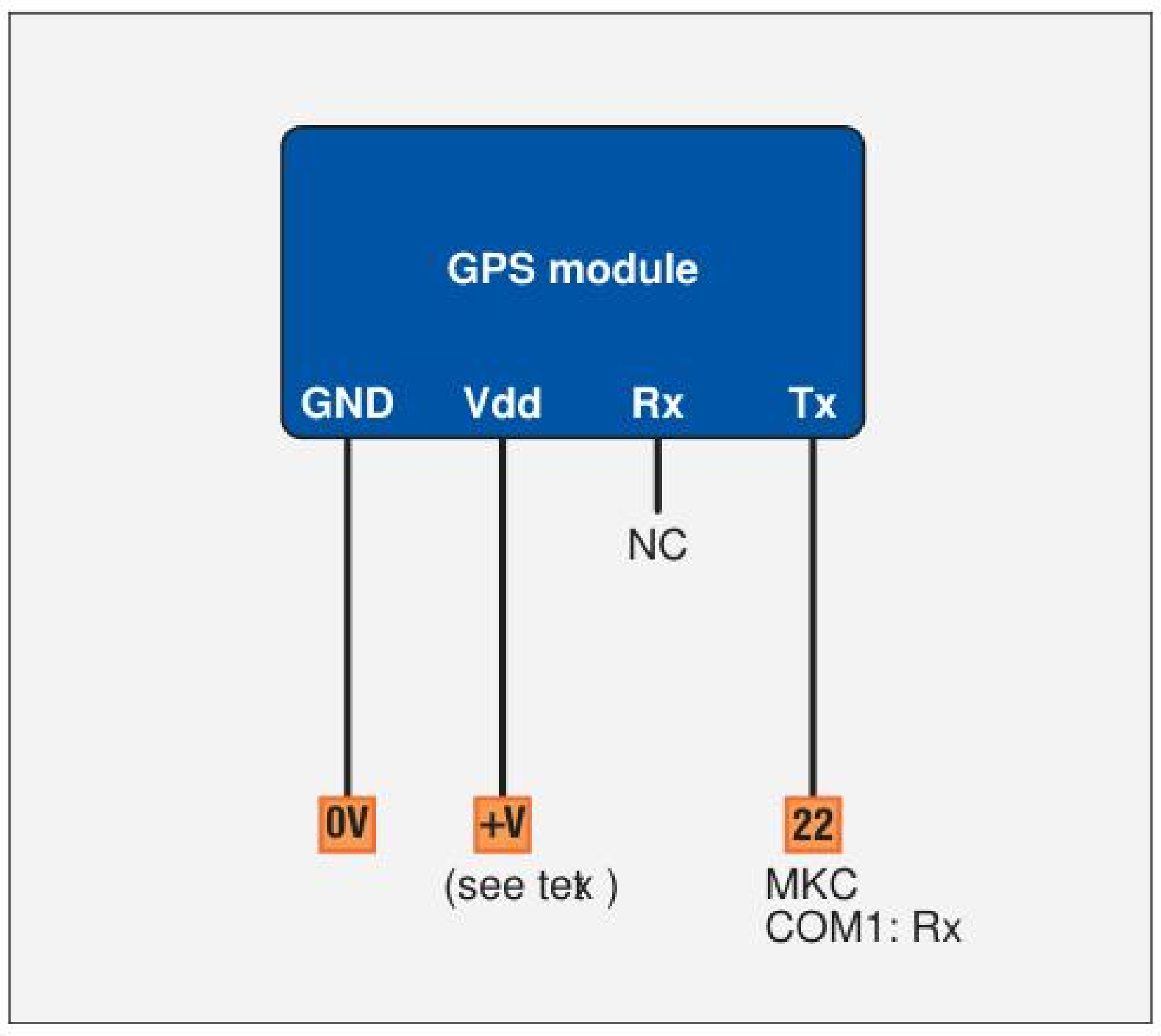


Fig.6. The circuit diagram for connecting a GPS module – only three connections are required to be made.

```
OPEN "COM1:9600" AS #1
```

This one line will configure pin 21 as the Tx pin, and pin 22 as the Rx pin (the two pins associated to COM1). However, we are not actually using pin 21 in this application since the Micromite is only being used here to 'listen' to data sent from the GPS module.

The `9600` figure defines the baud rate. This value is chosen as it is the rate which the GPS module uses (as per it's datasheet). We do not have to concern ourselves about setting the 8-N-1 parameters discussed above since the GPS module uses the same default values as MMBASIC. If it did use different values, then the relevant parameters would also be set in the above configuration command (refer to the *Micromite User Manual* for the specific details).

The `#1` is a reference to a file buffer which is inside the Micromite. Any data that is received from the GPS module will reside in this buffer, as we will now see.

Data that is sent by the GPS module to the Micromite temporarily sits in what is referred to as the 'UART receive buffer'. It sits there until the buffer is either full (at which point new data received overwrites the 'old' data), or until the data is read out from the buffer by using the MMBASIC command:

```
data$=INPUT$(n,#1)
```

What the above command does is to extract *n* character bytes from the UART receive buffer and load them into the string that we have named `data$`. With the data bytes loaded into this string, you can then perform various checks on the data in order to make your program code behave in certain ways – we will see how to do this very shortly.

GPS sentences

We have mentioned that our aim this month is to receive and display the data that is outputted from the GPS receiver.

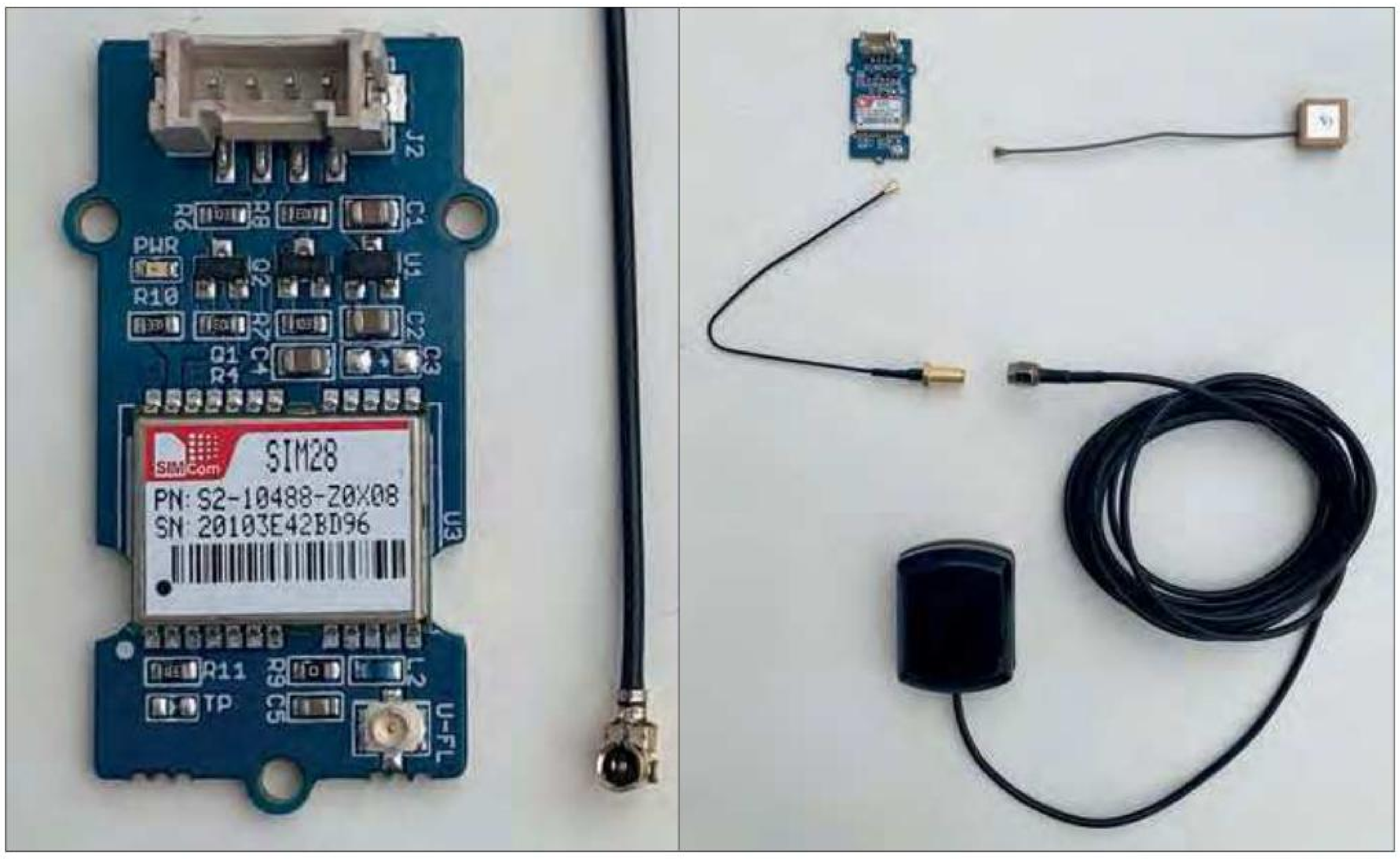


Fig.5. If required, use a uFL-SMA adaptor to allow a long-lead antenna to be plugged in instead of the smaller Grove antenna.

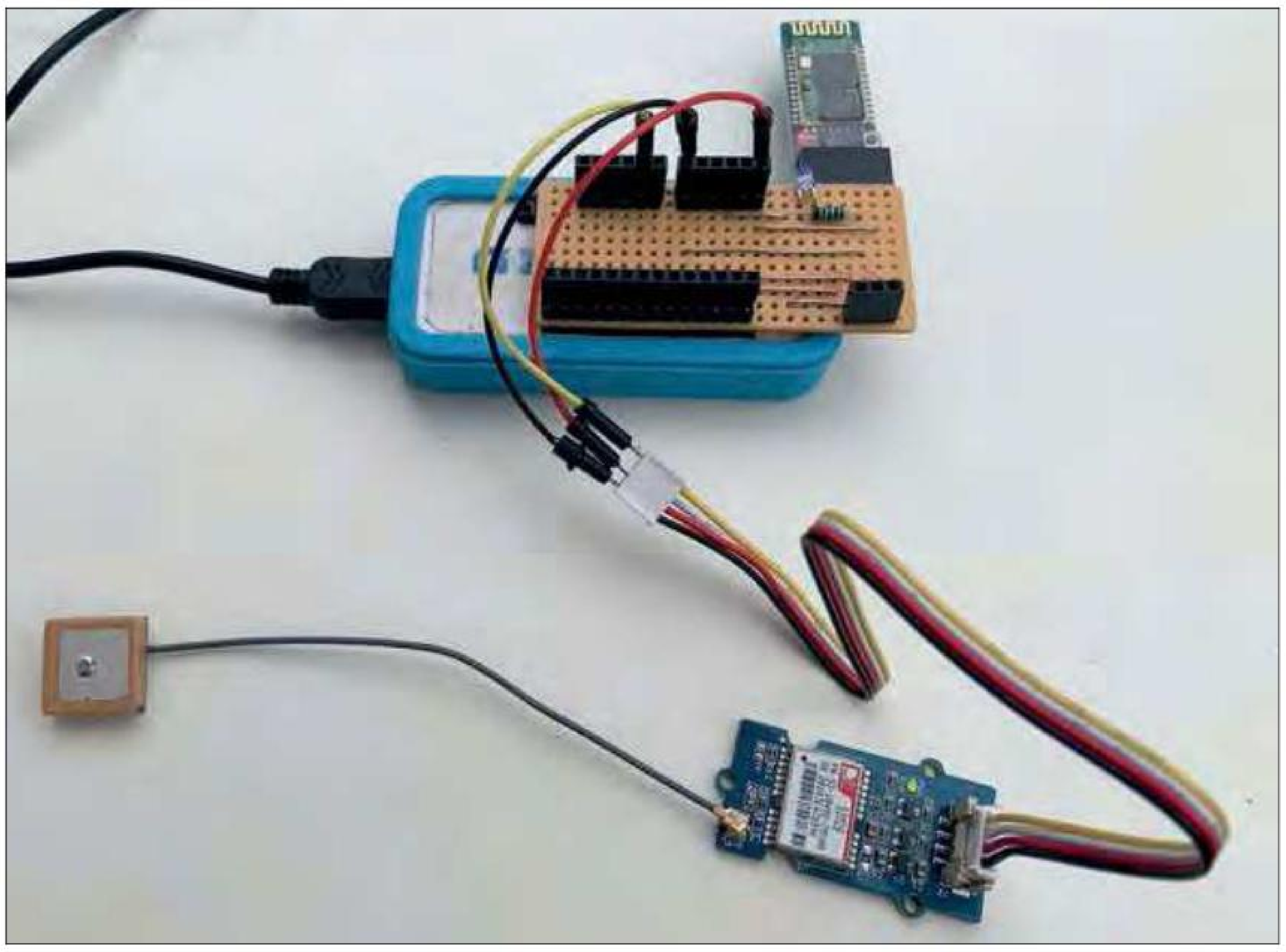


Fig.7. The Grove GPS module connected with jumper wires to the *MKC* and the Bluetooth module. (This is a 5V-powered module.)

The great thing about most GPS receivers (including the Grove GPS module used here) is that they continuously send out useful data such as current position co-ordinates, date, time, altitude, speed, number of satellites seen and so on. This data is typically sent every second on the GPS module's Tx pin. In fact, the various data elements are sent out in clearly defined 'sentences', the format of which was developed by the National Marine Electronics Association (which is why these sentences are better known as 'NMEA sentences').

Table 1 shows several of the different NMEA sentences that exist, and Table 2 shows all the elements that are contained in the specific sentence that we will be using (the GPRMC sentence). As can be seen, the GPRMC sentence contains coordinate information (longitude and latitude), and also the current date and time.

Note that each data element in an NMEA sentence is separated by a comma; and another bonus is that the data is outputted in the standard ASCII format so it is very easy to display the received data. Note too that each sentence begins with a leading $ character (at the start of data element 1), and ends with a checksum data element that starts with a * character. We will be using these characters to help detect the start and the end of a sentence.

Viewing GPS Sentences

Now that we have the GPS module connected to the *MKC*, and that we also have a basic understanding of the MMBASIC commands used to configure and extract data from the Micromite's UART receive buffer, let's load a short program that allows us to view the data from the GPS module. Go ahead and download **GPS_Sentences.txt** from the October 2020 page of the *PE* website and install it on your *MKC*. On running the program, you should see some text appear on your terminal screen similar to that shown in Fig.8. Note that when you first apply power to your GPS module, there may be a delay of up to two minutes before any data is displayed on the screen. This is purely because a GPS module needs to see and lock on to several GPS satellites before it can output any meaningful data. With the Grove GPS module, the start-up time is approximately 15 seconds, with coordinate data starting within 30 seconds.

Referring to the screenshot in Fig.8, you will see that there are four NMEA sentences displayed. The left-most (ie, first) data element of each line shows the type of NMEA sentence (refer to Table 1). If you look at the last line, then you can see that this is the GPRMC sentence, which is the one that contains all the data that we are interested in. It is shown in Fig.8 with a total of 10 received data elements, the definition of which are outlined in Table 2. So from the screenshot shown in Fig.8, and by referencing Table 2, we can extract the following useful data from the GPRMC sentence:

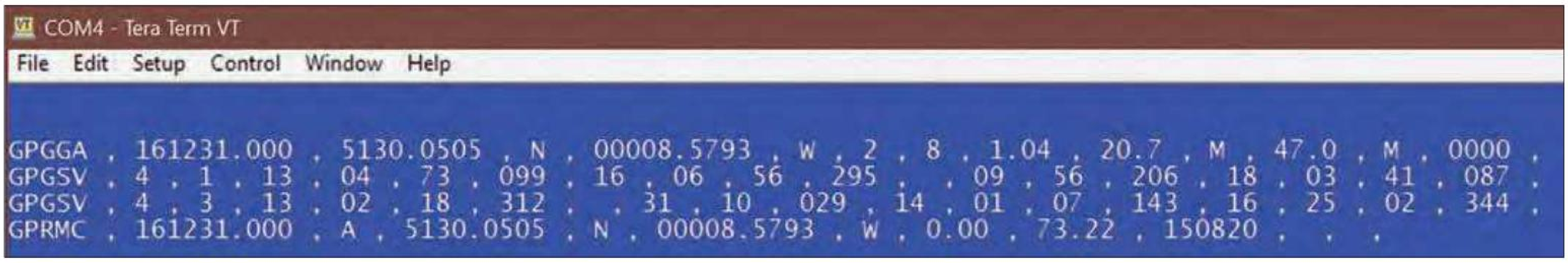
- Time = 161231.000 = 16:12:31
- Data is valid (Data element 3 = A)
- Latitude = 5130.0505 N

Table 1: Different types of NMEA sentences

NMEA sentence	Meaning
GPGGA	Global positioning system fix data (time, position, fix type data)
GPGLL	Geographic position, latitude, longitude
GPVTG	Course and speed information relative to ground
GPRMC	**Time, date, position, course and speed data**
GPGSA	GPS receiver operating mode, satellites used in the position solution, and DOP values
GPGSV	The number of GPS satellite in view, ID number, elevation, azimuth and SNR values
GPMSS	Signal-to-noise ratio, signal strength, frequency, and bit rate from a radio beacon receiver
GPTRF	Transit fix data
GPSTN	Multiple data ID
GPXTE	Cross track error, measured
GPZDA	Date and time (PPS timing message, synchronised to PPS)

Table 2: The format of the GPRMC sentence. It contains all the data elements we require, so this is the sentence we use. Here, the example GPRMC GPS sentence is:
$GPRMC, 161229.487, A, 3723.2475, N, 12158.3416, W, 0.13, 309.62, 120598, , A, *10

Date element	Example	Description
1. Message ID	$GPRMC	RMC Protocol Header
2. UTC time	**161229.487**	**hhmmss.sss**
3. Status	A	A = data valid or V = data not valid
4. Latitude	**3723.2475**	**ddmm.mmmm**
5. N/S indicator	N	N = north or S = south
6. Longitude	**12158.3416**	**dddmm.mmmm**
7. E/W indicator	W	E = east or W = west
8. Speed over ground	0.13	knots
9. Course over ground	309.62	degrees
10. Date	**120598**	**ddmmyy**
11. Magnetic variation		Degrees (E = east or W = west)
12. Mode	A	A = autonomous, D = DGPS, E = DR
13. Checksum	*10	
14. <CR><LF>		End of message termination



```
GPGGA , 161231.000 , 5130.0505 , N , 00008.5793 , W , 2 , 8 , 1.04 , 20.7 , M , 47.0 , M , 0000 ,
GPGSV , 4 , 1 , 13 , 04 , 73 , 099 , 16 , 06 , 56 , 295 ,  , 09 , 56 , 206 , 18 , 03 , 41 , 087 ,
GPGSV , 4 , 3 , 13 , 02 , 18 , 312 ,  , 31 , 10 , 029 , 14 , 01 , 07 , 143 , 16 , 25 , 02 , 344 ,
GPRMC , 161231.000 , A , 5130.0505 , N , 00008.5793 , W , 0.00 , 73.22 , 150820 ,  ,  ,
```

Fig.8. On running **GPS_Sentences.txt** you should see something similar to this, which indicates that GPS data is being received.

- Longitude = 00008.5793 W
- Date = 150820 = 15 August 2020

When the program is up and running, you will see a continuous stream of batches of NMEA sentences received from the GPS module (typically once every second). However, we are only interested in the GPRMC sentence so now make the modification to the program code that is *outlined in the detailed comments*. Once you have done this, re-run the program and you will now only see the GPRMC sentences. Note that this makes it easier to see the time, as the displayed GPRMC sentence is effectively updated each second.

We will not repeat how the code works here; instead, please do take the time to study the detailed comments in the program code.

GPS data conversion

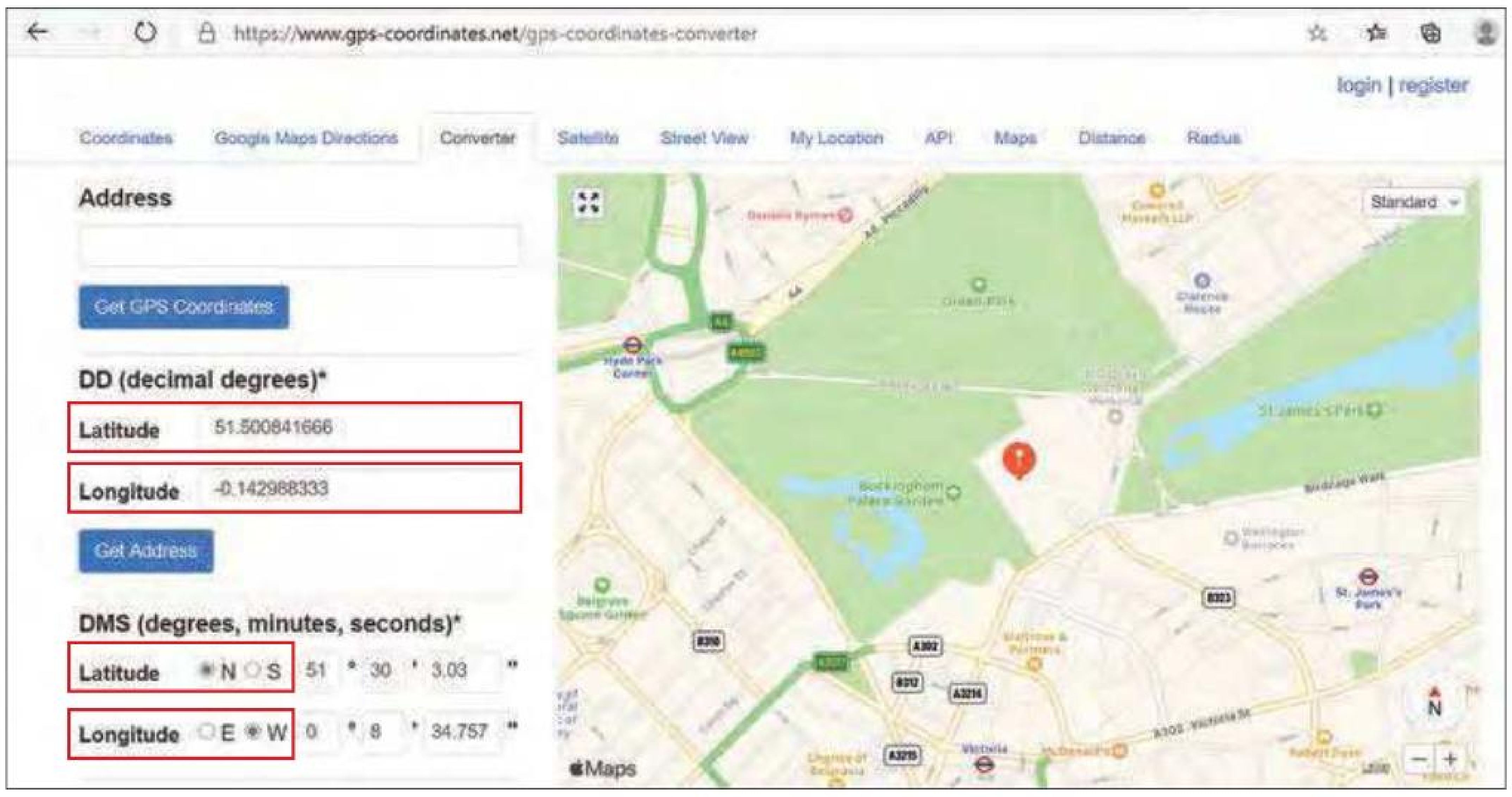
Data from a GPS receiver contains lots of information; however, some of the data is in a 'standardised' format and hence will need converting before use (using simple calculations). For example, the observant of you will have noticed that the GPS time is not correct. Instead it will be in a format referred to as 'Universal Time Coordinated' (UTC), so we will first need to add (or subtract) an offset in order to convert it into the true 'local time', not forgetting to take into account any daylight saving. At the time of writing, for UK readers this simply means adding on one hour to the GPS time.

The other data that will need converting are the GPS longitude and latitude coordinates. By converting them into a 'decimal degrees' format we will then be able to use the converted data to plot a location on a map (by manually entering the decimal degree coordinates into a website). Let's do this next.

Where are you?

To check the accuracy of your GPS receiver, we will now work through the steps of how to plot your position on a map using the GPS coordinates received by your module. Be sure to follow these steps precisely!

1. Run the existing program code to check that the third data element is shown as an A character (and not a V). If you see an A then move to step 2, otherwise wait for an A to appear (you may need to check the antenna is inserted correctly, and/or relocate nearer to a window to improve the GPS reception)
2. Make a written note of data elements 4-7 (ie, 5130.0505 , N , 00008.5793 , W)
3. Convert data element 4 (ie, 5130.0505) by removing the first two digits (51) and dividing the remaining digits by 60 (ie, 30.0505/60 = 0.500841666). Now add the first two digits back on to the result to yield: 51.500841666
4. Convert data element 6 (ie, 00008.5793) by removing the first three digits (000) and dividing the remaining digits by 60 (ie, 08.5793/60 = 0.142988333). Now add the first three digits back on to the result, producing: 0.142988333
5. Now visit the website: **https://www.gps-coordinates.net/gps-coordinates-converter** and scroll down the page to see the input boxes for DD (decimal degrees), as shown in Fig.9.
6. Enter the converted latitude value (51.500841666) and longitude value (0.142988333) calculated above
7. Look at data element 5 (N or S) and data element 7 (E or W) and ensure you select matching letters in the DMS section (underneath where you entered the DD coordinates)
8. Now zoom in on the map and it should show your current position with a very high degree of accuracy. If you are not seeing the correct location, then simply check the above calculations again, and ensure that the values have all been correctly entered.



Address

DD (decimal degrees)*

Latitude 51.500841666

Longitude -0.142988333

DMS (degrees, minutes, seconds)*

Latitude ● N ○ S 51 ° 30 ' 3.03 "

Longitude ○ E ● W 0 ° 8 ' 34.757 "

Fig.9. Visit **www.gps-coordinates.net/gps-coordinates-converter** – then scroll down the page and enter your Decimal Degrees values, along with selecting the N/S and E/W options. It will then plot your exact location on the map.

As a challenge, why not modify the program code to automatically perform the above conversion calculations on the coordinates and display those converted values on the terminal screen instead. Hint: use the `LEFT$`, `MID$` and `VAL` commands to manipulate the data.

A simple GPS Clock

Next, we will download a short program to display the time. There is nothing complicated about this code – it simply extracts the time (data element 2) from the GPRMC sentence, and then it uses ESC codes (see *Part 8*, *PE*, September 2019) to format the time on your terminal screen (ie, in the TeraTerm window).

Comments are present throughout the program code, and the file you need to download is **GPS_Clock.txt**. (from the October 2020 page of the *PE* website). Do be sure to look through the code to ensure you can see how the program works. As always, drop me an email if there is something you don't fully understand.

Next month

The topics covered in this article have set us up nicely for next month, when we will be showing you how to create a modern-looking analogue-style clock built from a ring of 60 digital RGB LEDs. Until then, have fun coding!

Questions? Please email Phil at:
contactus@micromite.org

Pedal Power Station!

Part 1

Words and design: Julian Edgar
Photos: Georgina Edgar

This project generates real power, keeps you fit and best of all, you can have fun building it from scrap and salvaged parts at low cost!

Here's a project that can develop lots of power – just from pedalling. How much? Try 25A at 13V (that's 325W)! Use it to charge a 12V battery, or via a normal car adaptor, any device that uses USB 5V charging.

Alternatively, you can easily configure the project as an exercise bike, adding a low-cost display that shows precisely how many watts of power you are developing. To make it even more exciting, you can add high-power lights that glow brightly with your pedalling effort – so much better than just dissipating your energy into a friction brake. Or, if you live in a hot climate, you can even power a large fan to keep you cool as you exercise!

There are three things that make this project better than most pedal-based generators of the past:

1. An efficient two-stage gearing system
2. A large generator
3. Low-cost prebuilt electronic modules to condition the output and display power.

This issue covers the bike mechanicals and the generator, while next month we'll explain the electronic instruments you can fit, and how the electrical output can be conditioned.

However, note that the series is presented here more as an 'ideas forum' rather than a step-by-step project. Why? Because it depends on what you can scrounge for nearly nothing. To buy all the parts brand new would be quite expensive – but, by using other people's discarded items it really can be a cheap project.

First, let's take a look at the main components.

Choosing a generator

The suggested generator is a discarded or low-cost DC low-voltage motor. As many of you will know, when a DC brushed motor has its shaft driven, it becomes a DC generator. And, since motors are much more common than generators, looking for discarded motors is the first step in sourcing a generator suitable for pedal power.

Almost any low-voltage brushed DC motor can be used, but when picking one look for a design that:

- Is large and heavy, with an output shaft that is at least 8mm in diameter
- Uses thick cables, indicative of a high-current design.

When you think you may have found a suitable generator, it's easy to test it. Connect a load to the generator (eg, a 50W incandescent car headlight bulb) and then use your electric power drill to spin the generator shaft. (To drive the generator, just lock the generator's shaft in the drill's chuck.) Monitor the output voltage with a multimeter and ensure you don't spin the generator so fast that the bulb's voltage rating is exceeded. (If it looks like the voltage will rise too high, add another parallel bulb – the greater the load, the lower the output voltage.) If the 50W bulb does not light, try a lower power one. If the 50W bulb glows brightly when powered by the drill, it's likely that the generator will be suitable for pedal power.

Another test is to short the generator's output (that is, connect its wires together) and then turn the generator by hand. The resistance to turning should be much higher than when the generator is open circuit.

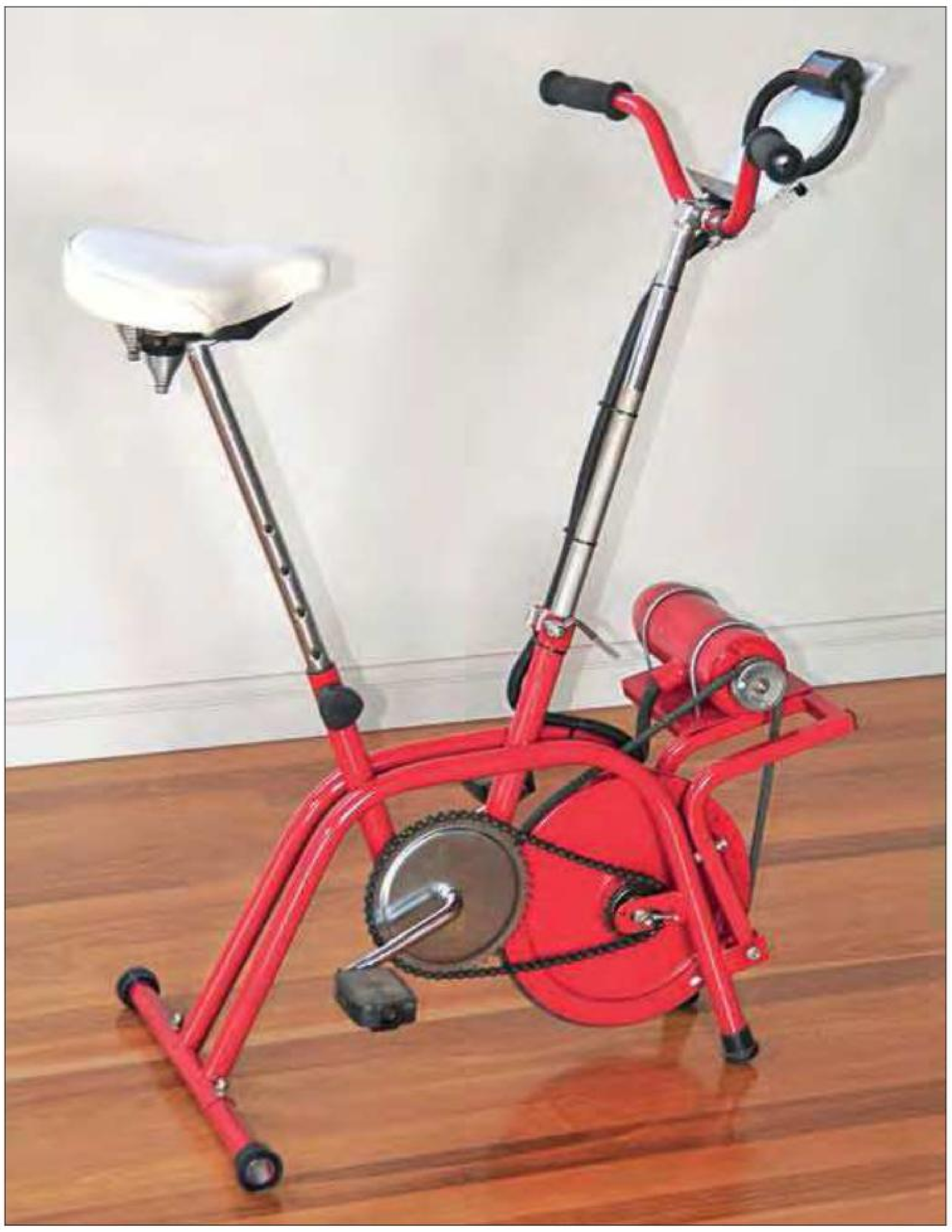


Fig.1. The *Pedal Power Station* in its completed form. A powerful DC generator is driven by two stages of step-up gearing, giving a good output at a normal pedalling speed. The prototype can develop a peak of 350W – and probably more with a stronger rider than me!

Use a stepper motor?

Instead of using a DC brushed motor as the generator, a large salvaged stepper motor can be used instead. This has two advantages: stepper motors used as generators (alternators, really) typically have a high output at low rotational speeds, and these motors are usually equipped with ball bearings rather than the plain bushes used in many DC brushed motors.

However, there are two downsides – the wiring is a little more complicated, and large stepper motors are more difficult to find as low-cost discards. (Of course, you can use a small stepper motor, but the power output will be much lower.)

As with DC motors, the easiest way of assessing the suitability of a stepper motor is to drive it with an electric drill and see how well it powers a load. But how do you do the wiring?

Most commonly found steppers are 6-wire designs, with, as Fig.2a shows, two electrically separate centre-tapped windings. Use a multimeter to measure the resistances of the coils until you ascertain which wires are which. Then place an incandescent light (eg, a 20W 12V bulb) across a winding (for example, connections 1 and 2 in Fig.2a) and spin the stepper motor with the drill. The lamp should light brightly on the AC waveform generated by the one phase.

Next, short those two wires together. The stepper should now be now much harder to turn, with a distinct gritty 'cogging' action.

If the stepper motor performs well in these two tests, it should be suitable for pedal-generator use.

To gain a DC output, wire diodes as shown in Fig.2b. Specify the diodes based on the current you're able to generate from each phase – for example, you may need four 6A10 (6A) diodes – cheaply available online.

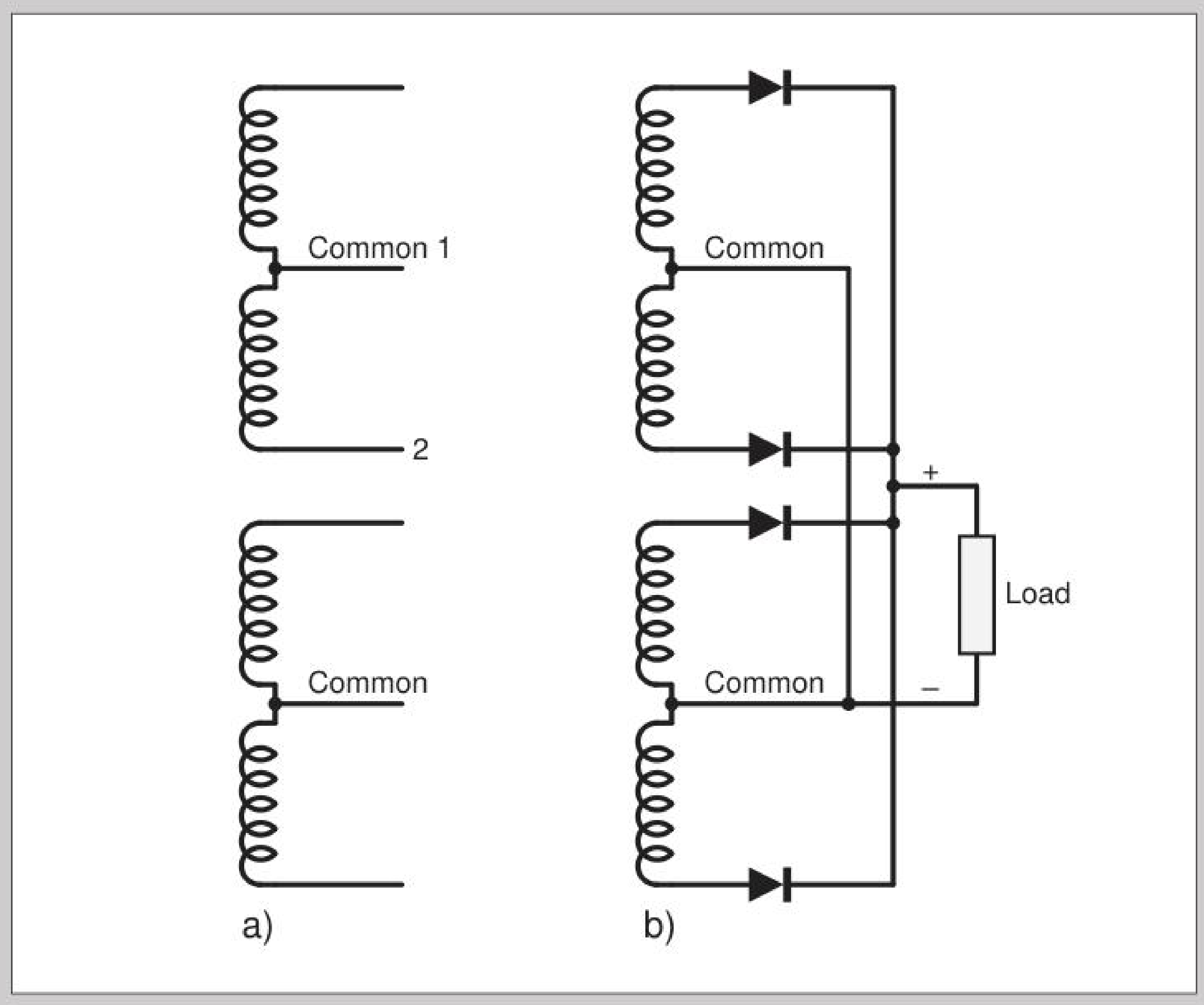


Fig.2a. Most salvaged stepper motors are 6-wire designs configured with two centre-tapped windings. Use your multimeter to sort out which wires are which.

Fig.2b. To gain a DC output, wire four diodes as shown here. Rate the diodes according to the current that you measured when testing each phase.

The starting point for my *Pedal Power Station* came when I found a discarded electric outboard motor. Often used on lakes that prohibit internal combustion engines, these motors are usually conventional brushed DC designs. (However, one wrinkle is that they may include speed control electronics within the motor housing. This prevents their use as a generator (the electronics don't work backwards!), but the speed control electronics can be easily bypassed just by connecting the output wires straight to the brushes.)

In addition to electric outboard motors, car starter motors that do not use reduction gears, wind generators and even the DC motors from old computer tape drives and similar can all be used. For a lower powered generator, a brushed motor from an electric scooter is cheap and readily available. A car alternator initially looks attractive, but unless you use really high step-up gearing, the alternator will usually have too low an output.

Incidentally, with second-hand motors, it's a good idea to open them up and give them an internal clean. (As the brushes wear, carbon gets distributed through out the motor internals.) While the motor is apart, grease the bearings (usually just plain bronze bushes) and, if the commutator is scored, smooth it with some fine emery paper, using a long strip wrapped around the commutator and a back and forth sawing motion. When re-assembling the generator, be careful to ensure that the carbon brushes go back properly into place – sometimes you'll need to juggle them a little to get them seated again on the commutator.

Fig.3. An old electric outboard motor makes an ideal pedal generator. However, any heavy-duty, low-voltage, brushed DC motor can be used. This electric outboard is a little smaller than the final one used on the *Pedal Power Station*.

Fig.4. The armature of the electric outboard motor used in the *Pedal Power Station*. When using a salvaged motor as a generator, it's a good idea to first disassemble it, lubricate the bearings and clean the commutator.

Pedal machine

The best starting point for the ‘pedal’ end of things is an old exercise bike. (Starting with a conventional bicycle frame will mean a lot more work.) Try to find an unwanted exercise bike with these features:

- Sturdy with a heavy-gauge steel frame
- Easy access to a driven wheel so that it can be adapted to drive the generator
- High gearing (so that the driven wheel rotates fast)
- As heavy a driven wheel as possible
- A driven wheel with either an inflatable tyre or alternatively, a flat edge to the rim
- Smooth operation – the bearings in good condition

Let’s take a look at these in turn. You want an exercise bike with a strong steel frame because you will need to either bolt or weld an additional frame to it, one that will support the generator. Light-gauge material will not have sufficient strength to support this attachment. In addition, the design of the bike needs to be such that there is clearance (usually in front of the wheel) for the generator to be located.

The higher the pedal gearing, the faster the driven wheel will rotate – and so, when you adapt a generator to it, the faster the generator will turn. Typically, the generator will have a higher output at higher speed – so faster equals better! And why a heavy driven wheel? A heavy wheel act as a flywheel, helping to smooth your pedalling power inputs.

The exercise bike shown below was bought from the local recycling centre for £10.

If the exercise bike wheel has an inflatable tyre on it, the easiest way to drive the generator is to install a small-diameter rubber-tyred wheel on the generator’s input shaft and have this press against the tread of the bike’s tyre. This rubber-to-rubber contact will transmit a lot of power without slipping, and the inflated tyre on the bike will cushion the contact, giving good drive even if the two wheels aren’t perfectly round.

If the exercise bike has a smooth and flat edge to the rim, you can drive the generator via a belt. Two different types of belt are commonly available: the classic ‘V’ belt and the more modern multi-rib belt. (If you lift the bonnet of your car, you will see one or other of these belt types.) While not designed to do this, a large flat wheel on an exercise bike will drive either of these belt designs without slipping. However, on the generator, it is important that you install a pulley of the appropriate design – that is, either a V-belt pulley or a multi-rib pulley.

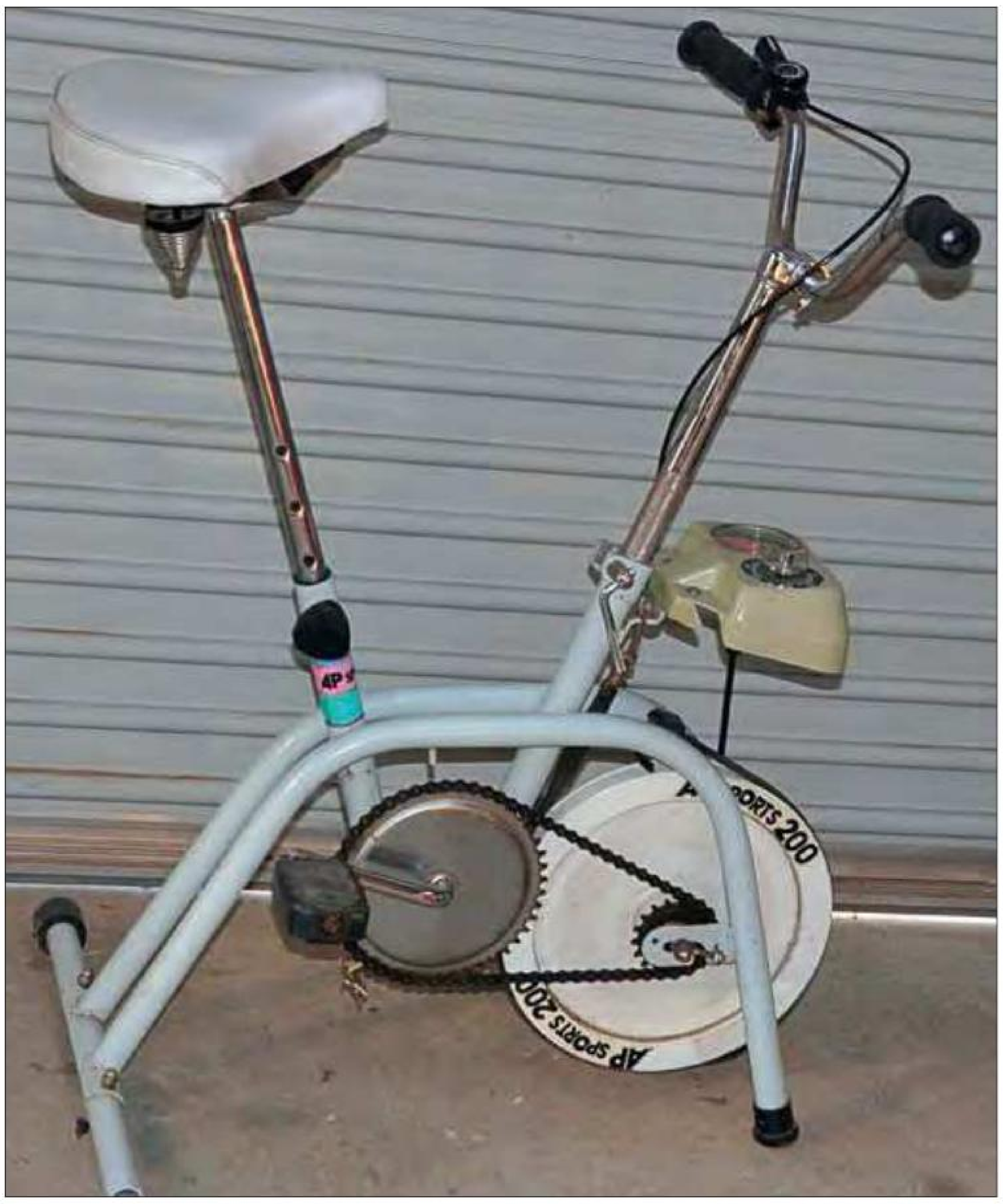


Fig.5. Starting point for the *Pedal Power Station* was this old exercise bike that was bought from a local recycling centre for £10. Points to look for when selecting a suitable bike are a sturdy steel frame, a heavy driven wheel, high step-up gearing and space in front of the driven wheel to mount a generator. Also check that the wheel and pedal bearings rotate freely.

Gearing

The *Pedal Power Station* pictured below uses 16.2:1 step-up gearing. The standard chain drive stage gives a step-up of 2.8:1, and the second belt stage, 5.8:1. At a cadence (pedalling speed) of 90 rpm, this gives a generator speed of about 1450 rpm, a speed which was confirmed using a laser tachometer.

Fitting the pulley

Of all the steps in making the *Pedal Power Station*, installing a pulley (or rubber-tyred wheel) on the generator shaft is likely to be the trickiest – so let’s take a look at that.

Depending on how the wheel or pulley is designed, attachment to the shaft might be via a press-fit or slip-on with a grub screw. In either case, the opening in the wheel or pulley is likely to be larger than the diameter of the generator shaft, so requiring a reduction sleeve. It is important that the sleeve is concentric and that its inner and outer diameters are precisely the correct sizes. Unless you have a metal-working lathe, this might be one job to take to a

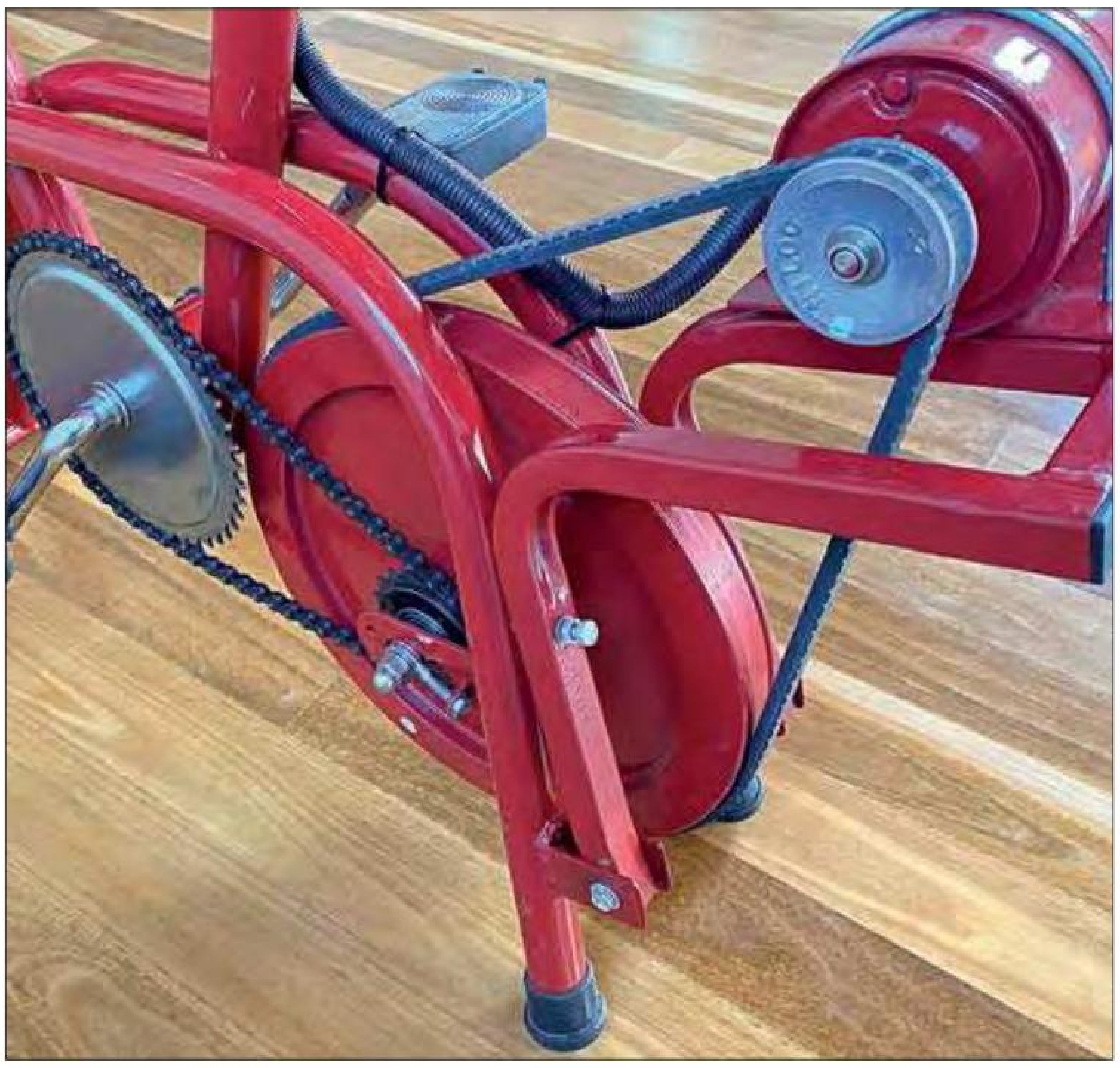


Fig.6. Here the generator is driven via a V-belt. (Alternatives are to use a direct friction drive or a multi-rib belt.) The V-belt runs on the flat surface of the original bike flywheel and drives a new pulley that has been fitted to the generator. The generator is mounted on a welded steel tube frame that attaches to the exercise bike. Note the lower pivot points and the upper bolts that allow the belt tension to be adjusted. The generator’s steel frame could also have been bolted together or made from wood.

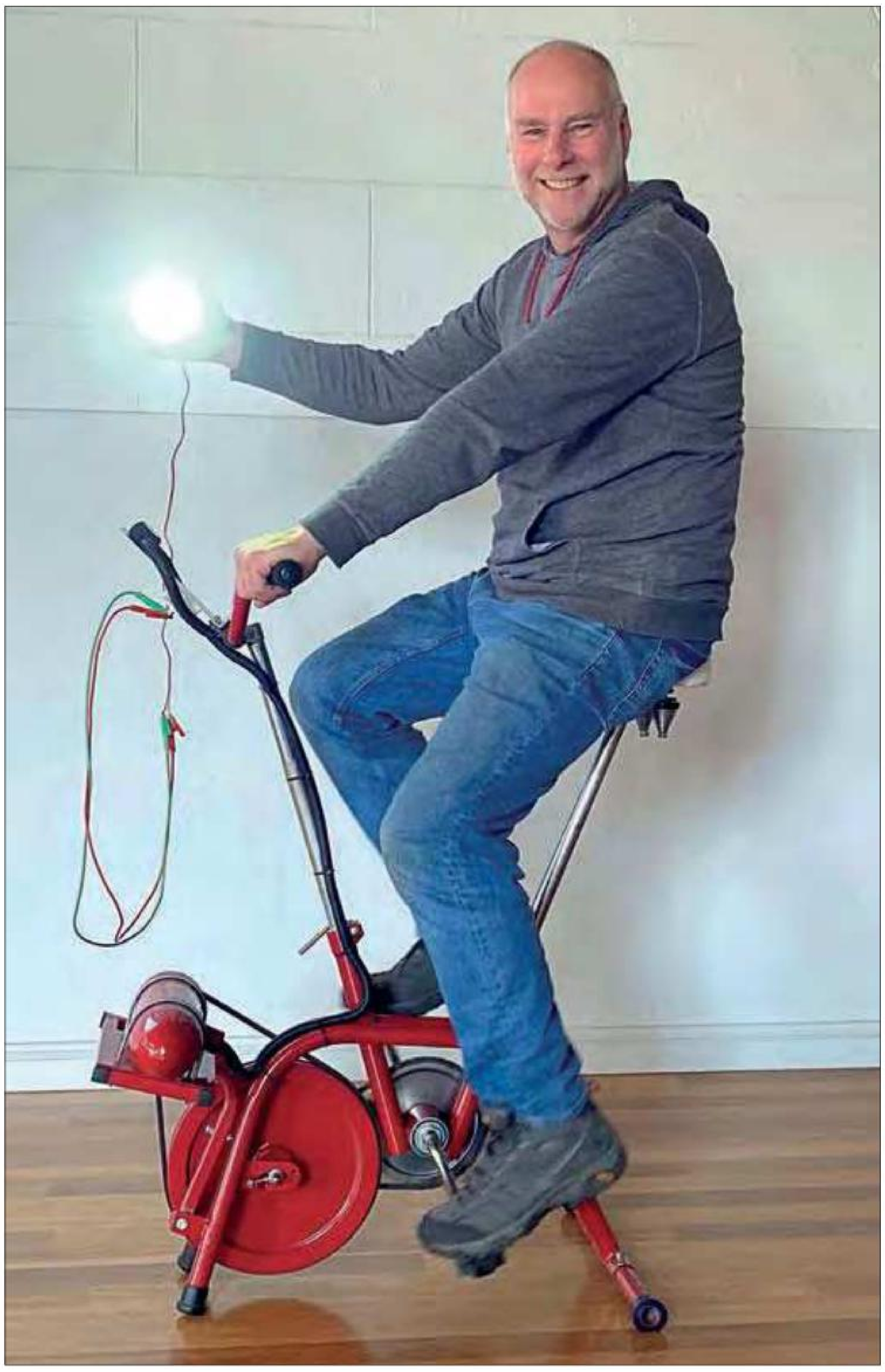
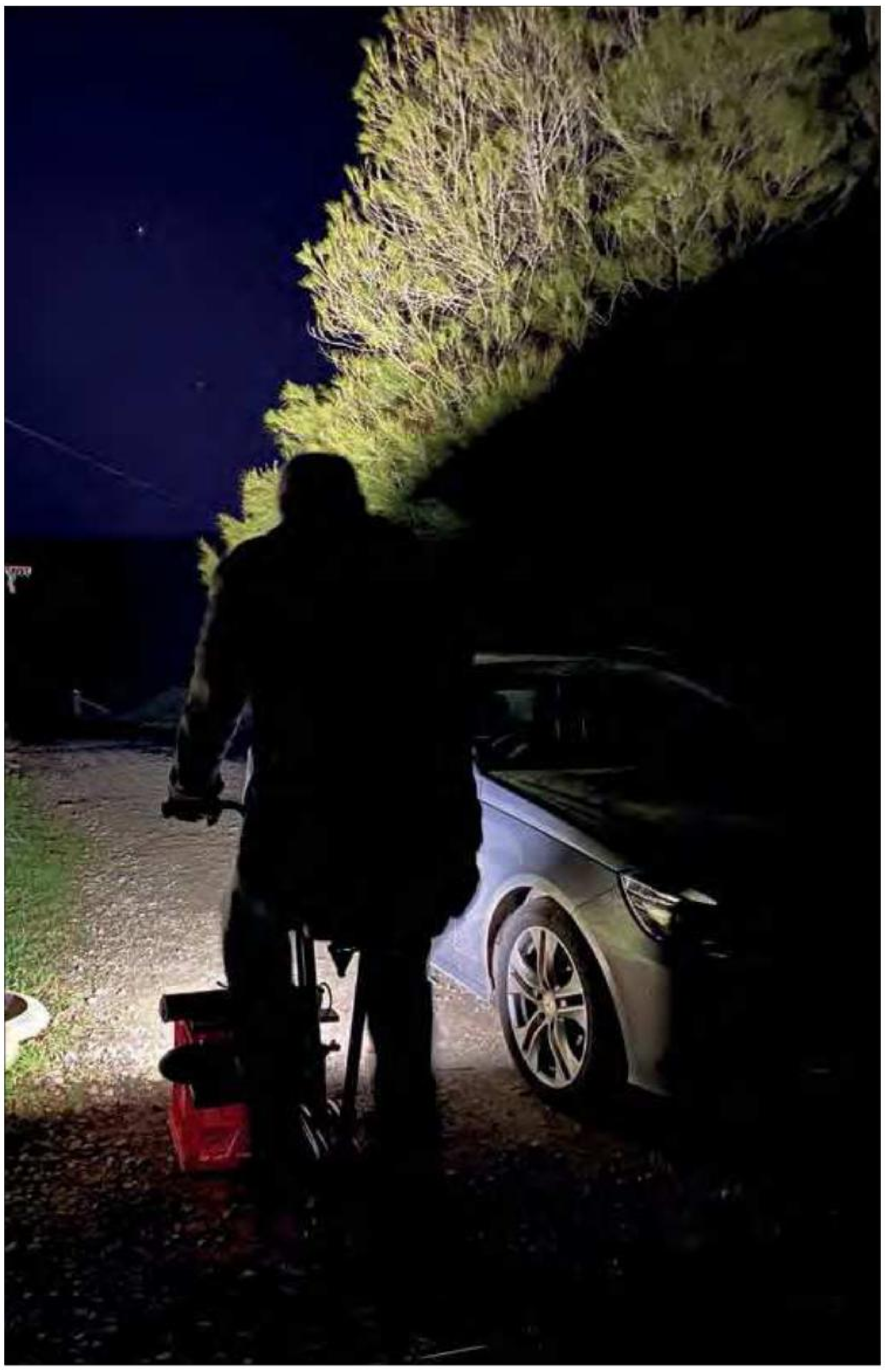


Fig.7. The *Pedal Power Station* is powerful enough to run almost any 12V LED light source – from one as modest as this 10W LED…



…to a pair of 90W LED light bars. And yes, without the pedal-powered lights, this view is pitch black!

small machine shop and ask for an appropriate bush to be made. (Don't forget to take the generator and pulley or wheel along with you.)

If you're happy with a rough-and-ready approach, you may be able to use a short length of rubber hose to make the bush. The rubber can be stretched over the generator shaft (ie, it doesn't have to be precisely the right size) and compressed within the wheel or pulley opening. The rubber is also easily sanded or filed to size. If taking this approach, ensure that the pulley or wheel cannot come loose – the last thing you want is it flying off at high speed!

Flywheels

When you pedal, you don't develop a smooth torque output. Instead, at the top and bottom of the pedal stroke, the torque greatly decreases as the legs pass Top and Bottom Dead Centres. This is not noticeable on a bicycle being ridden on the road, because the inertia of the bike smooths these variations. However, with a stationary pedal generator the result can be a jerky development of power.

The exercise bike shown here came with a heavy steel flywheel as standard, helping smooth the flow of power. However, many exercise bikes don't take this approach. While I haven't tried it, some on-line searches show people making flywheels by filling small bike wheels with suitably reinforced concrete. Heavy steel weights bolted symmetrically to the rim should also work well. In all cases, ensure that the result is structurally strong enough to withstand the centrifugal forces without failure. If there is any uncertainty as to this, enclose the wheel in a guard.

The cheapest way of obtaining either a V-belt or multi-rib pulley is at a car dismantling yard or an auto electrician. At the latter, you'll normally find plenty of alternators being thrown away – complete with their pulleys. If you are using a small rubber-tyred wheel, appropriate wheels are used on a wide range of goods that are often discarded. (Or of course you could also – gasp – buy a new one!) In both cases, select the smallest available wheel or pulley for the generator. This gives you the greatest step-up in gearing.

Once you have a pulley or wheel installed on the generator, you can work out how the generator is to be mounted. Let's take the 'belt-drive' approach first.

Position the generator in front of the exercise bike wheel and, using a piece of wire looped around the pulley and wheel, measure how long the required belt needs to be. Take the piece of wire to an auto parts store and buy the appropriately sized belt – both V and multi-rib belts are available in an enormous range of lengths.

If you are using a rubber-tyred wheel, push the wheel up against the exercise bike's wheel and look at where the generator will need to be positioned.

When positioning the generator, ensure that there is still room for the rider to pedal – be especially careful to give the required toe clearance.

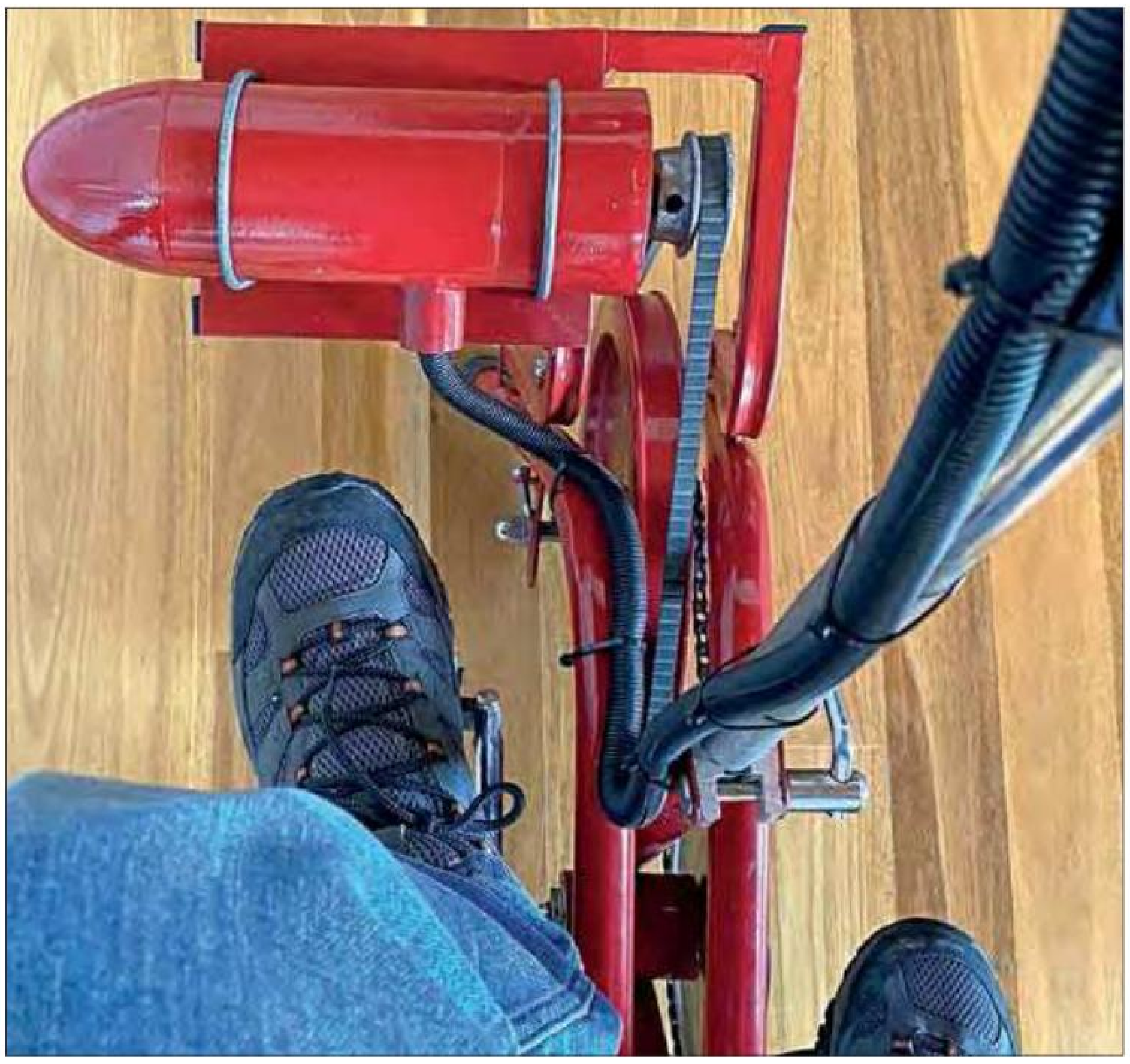


Fig.8. When locating the generator, be careful to retain sufficient toe clearance during pedalling.

Frame

You will then need to make a frame that attaches to the exercise bike and holds the generator in its required position. This frame should be strong and rigid, both because the generator is likely to be heavy and because the generator needs to maintain its registration with the drive wheel.

This frame can be made in a wide variety of ways. For example, you could use square steel tube bolted together with angle brackets, or even make a frame from wood, screwed and glued together. The attachment to the exercise bike frame can be via bolts passing through holes drilled in the frame, or you can use U-bolts that pass around the frame members.

Having access to a MIG welder, I made the frame from 20mm square steel tube (1.6mm wall thickness), welded together. I then welded tabs to the exercise bike frame to allow attachment of the generator frame.

The way that the generator is attached to its supporting frame depends on the generator and the original design of its mounts. In my case, where the outboard motor had only a single mount that attached to a long pole, I cut the pole off and used U-bolts to hold the generator to a flat plate on the new frame.

Make the generator frame adjustable so that it can be moved slightly to either tension the belt or vary the pressure with which the rubber-tyred wheel pushes against the exercise bike wheel. Note that this adjustment mechanism doesn't need to be elaborate – eg, some washers under the bolts can be used to adjust clearances, or the generator frame can be hinged and a spring used to give the required force. Having built alternator mounts for cars, where the belt tension needs to be very high, I was surprised at how little tension was needed to transmit pedal power to the generator without slippage.

Whichever drive approach you are taking, the frame must position the generator drive directly in-line with the bike wheel. If this is done, and the generator is held rigidly in position, the belt will track truly on the flat wheel and the driven rubber wheel will not 'walk' off the bike wheel.

Next month

That's all for this month – next issue we will look at the different loads than can be driven by the *Pedal Power Station* and how you can fit a low-cost digital instrument to measure your power output.

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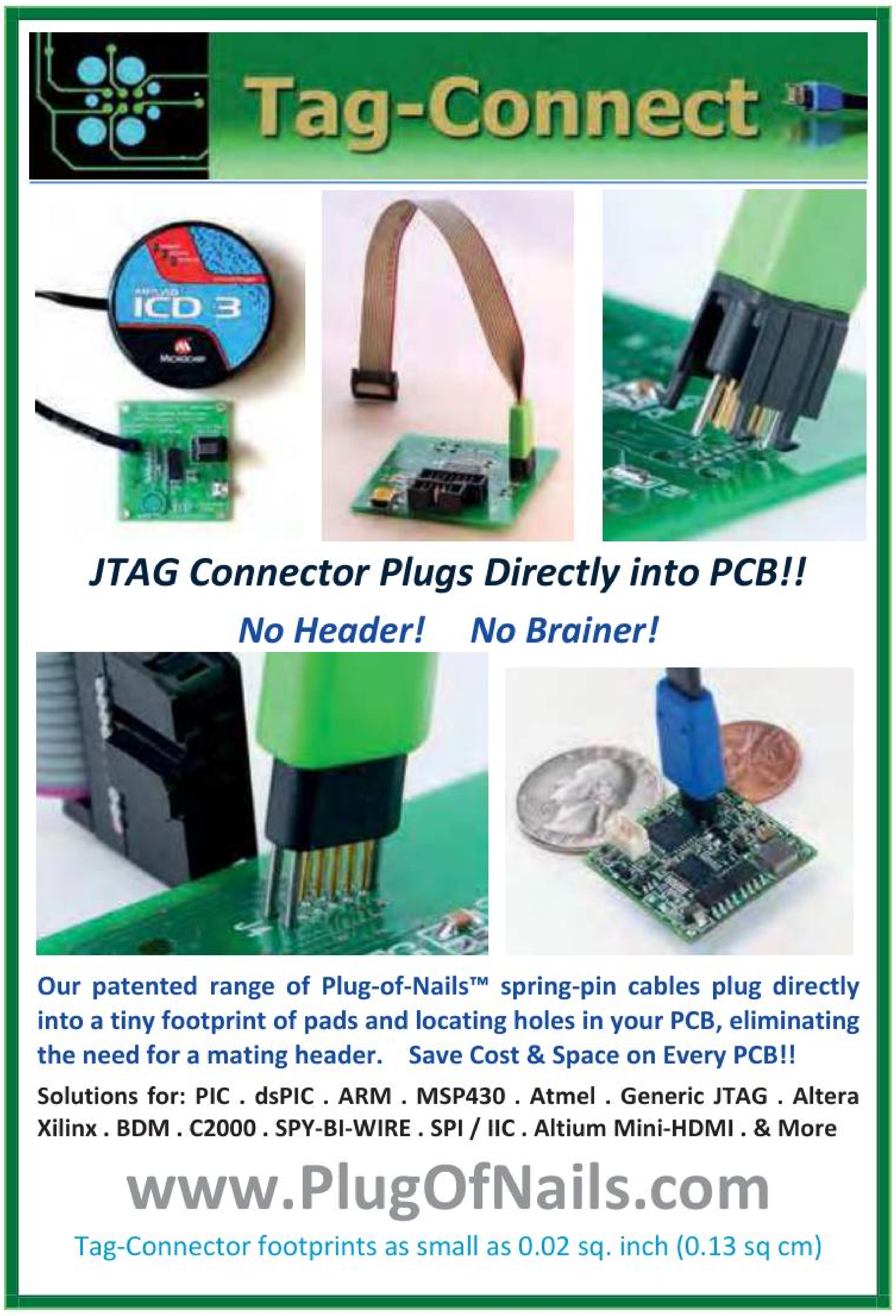

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Jim Gatenby

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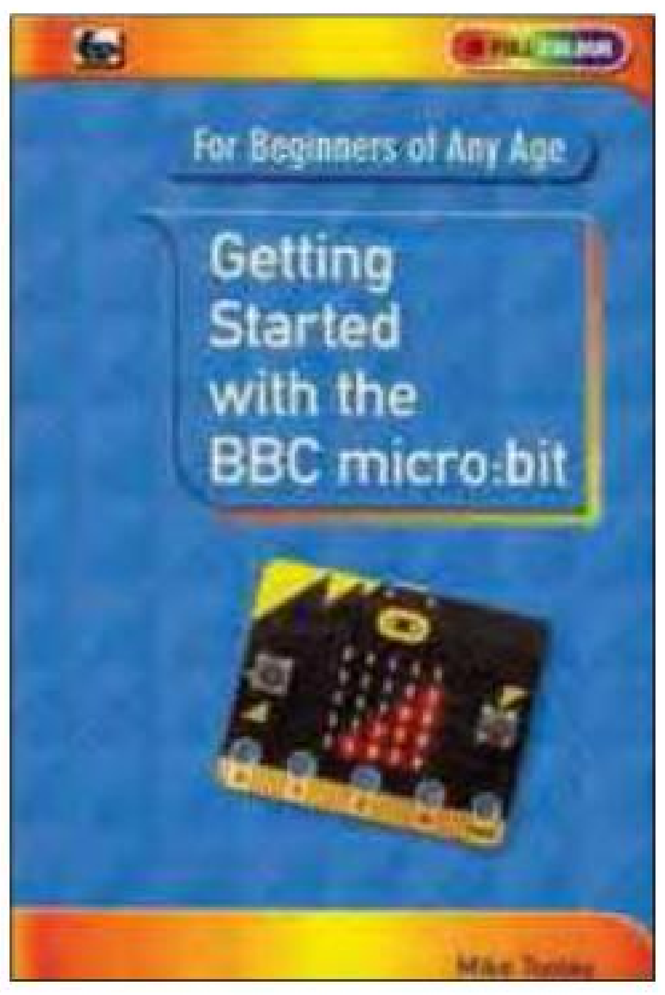

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Mike Tooley

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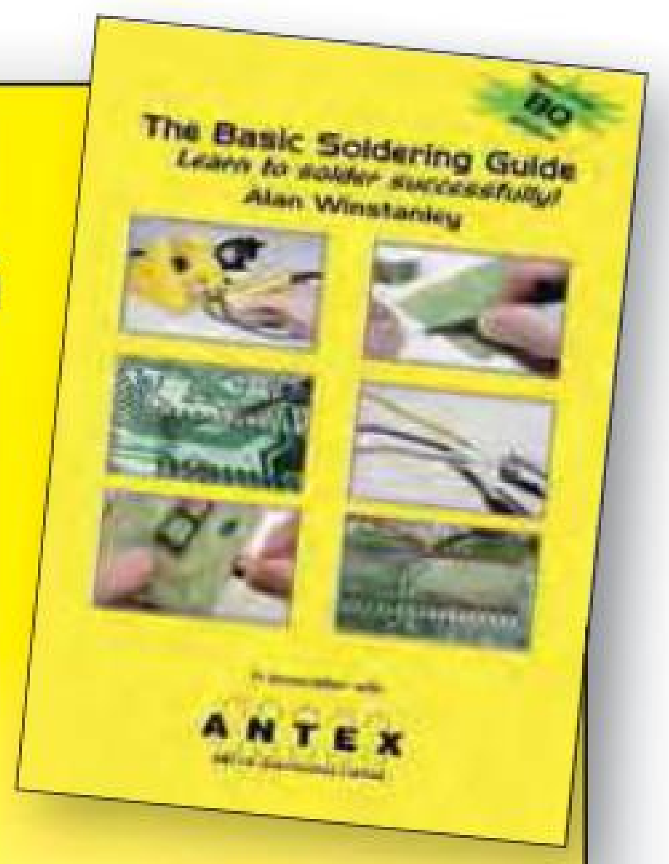

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AUDIO OUT

By Jake Rothman

Low-noise germanium transistor power supply

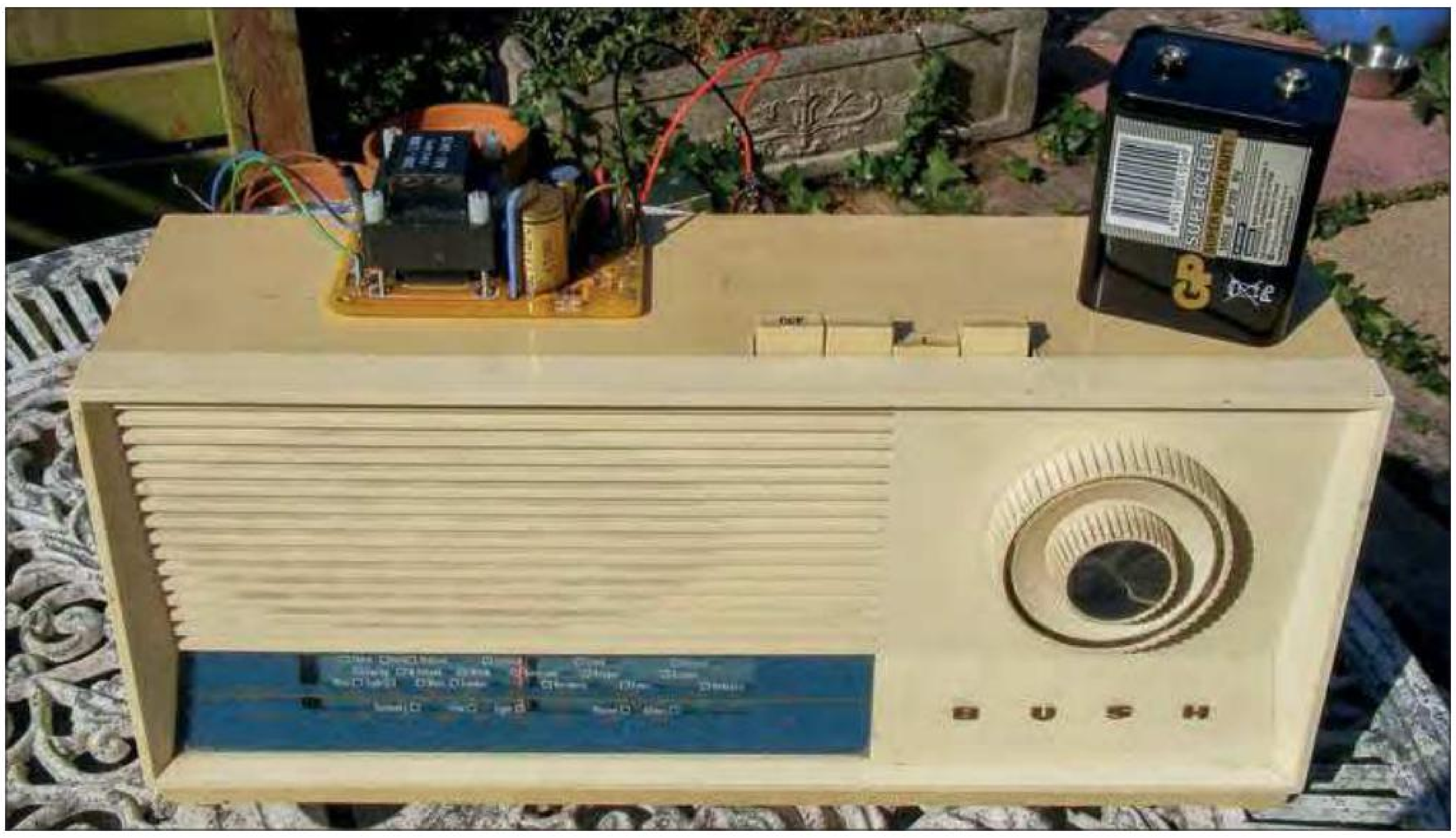

Fig.1. A Bush TR91 radio from 1961, a year older than me. It has seven PNP germanium transistors and is positive earth. The power supply replaced the expensive (£4.45) PP9 battery. It's still going strong, beautifully playing the *Shipping Forecast*.

Last month's theremin power supply board can easily be adapted to make a power unit for old germanium technology, such as Colorsound fuzz pedals, Deacy guitar amps and 1960s Bush radios (illustrated in Fig.1). These were described in my *Ge-mania* series of articles in *EPE*/*PE* from April 2015.

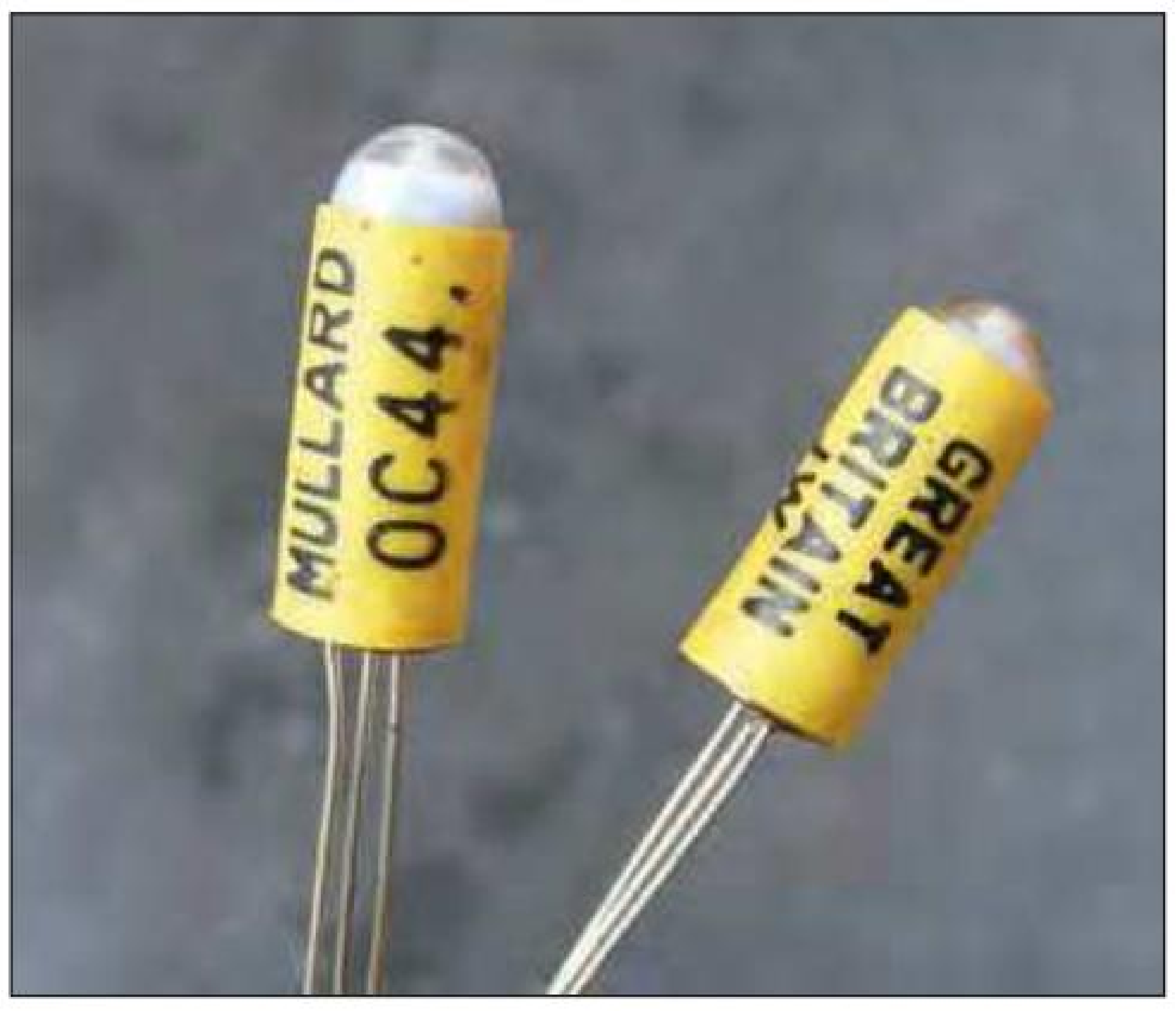

Fig.2. The Mullard OC44, regarded as the best transistor for fuzz boxes and also used in the mixer oscillator stage of the above radio. These are now an expensive antique, treat them to a decent power supply and remember to get the polarity right. The Texas 2G306 and Toshiba 2SA12,44,53 are good alternatives.

A good 95% of germanium transistors are PNP types, such as the OC44 shown in Fig.2; thus the circuits are normally *positive* earth with a *negative* supply rail. Most circuits are also 9V battery powered with a low current consumption of around 120mA max. Germanium circuits were generally simple due to the relatively high cost of early transistors, which resulted in poor power-supply rejection ratio. This means a low-noise power supply is essential.

Positive earthing

Setting up the LM317 voltage regulator to give *plus* 9V was described last month. One way to get a *negative* 9V rail would be to use the LM337 negative regulator. This would require a PCB redesign changing the regulator's pin-out and reversing all the polarised components. Assuming the power supply is only going to power positive-earth devices, a much simpler method is to connect the mains earth to the 9V positive output, redefining this rail as 0V, as shown in Fig.3. The old negative rail now becomes *minus* 9V and the positive rail is earth. This is akin to changing the battery round in a car to make it positive earth rather than negative. Some early (1964) VW Beetles had 6V positive-earth electrics, where the metal work was positive.

Noise reduction

Last month, we talked about using capacitors with a damping resistance (a Zobel network) to reduce rectifier-induced ringing and spikes. These occur when the transformer is unloaded for the period during which the rectifier diodes are turned off, and is thus free to ring for this period. The old Roberts RM30 radio in Fig.4 damped the ringing by having an extra transformer winding supplying a 6V filament lamp for illuminating the tuning scale. A nice 1970s trick of providing two solutions with one addition.

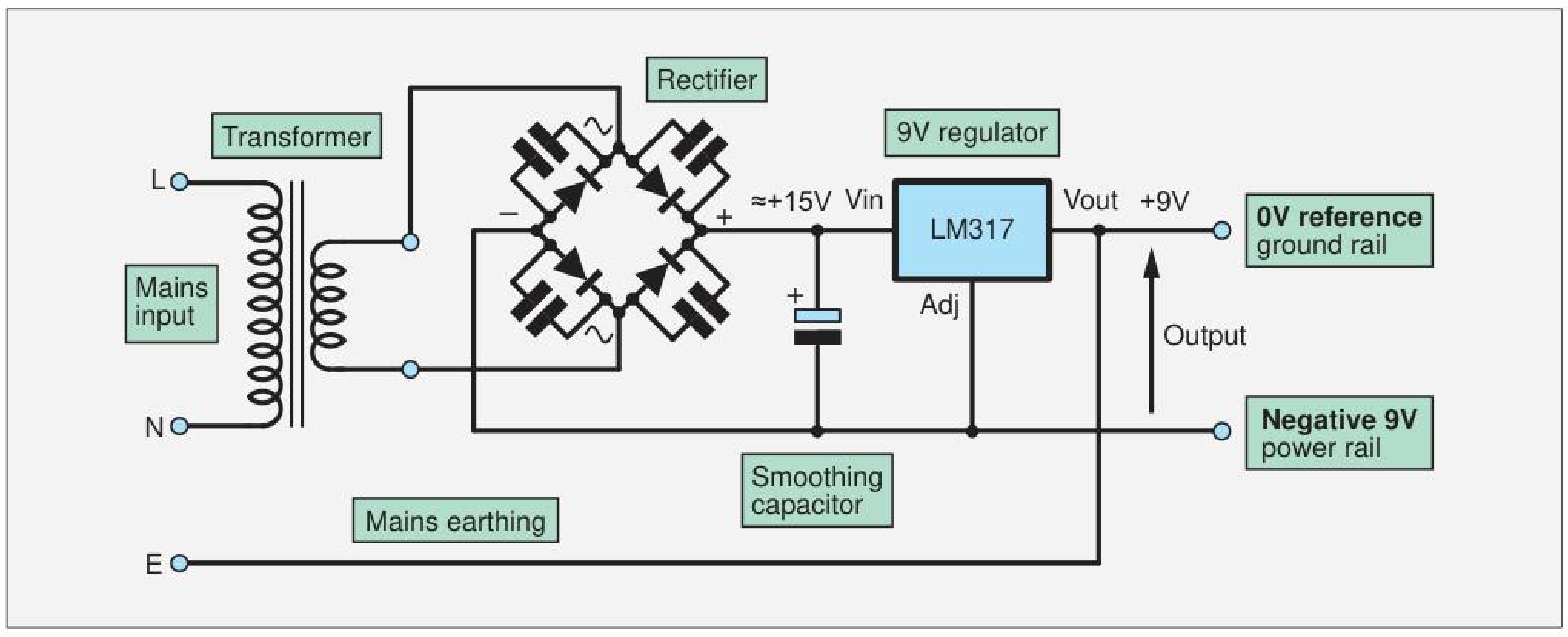

Fig.3. Redefining the 0V rail: connect the regulator's positive output terminal to mains earth.

ROBERTS RM30
VOLUME
LW
MW
VHF
TUNING
MAX

Fig.4. Old Roberts RM30 radio from the 70s, still going strong. In this example the tuning scale lamp damps the mains transformer. I only noticed this when the PSU hash level rose when the bulb failed.

The surprising benefit of significant capacitor equivalent series resistance (ESR) for regulator transient load stability was also described last month. I saw the effects of this when an old 1980s Harrison mixing desk was 'recapped' by a self-proclaimed audiophile 'guru' at huge expense. The replacement of numerous tantalum capacitors by low-ESR electrolytic types resulted in a myriad of instabilities.

Circuit overview

The power supply circuit we will build is shown in Fig.5. The main difference with the versions discussed last month is that the mains earth link has been connected to the positive output pin of the regulator, and not to what would normally be the 0V rail. Electrically speaking, it is still just a voltage source – but reversed relative to '0V'.

The Zobel network (Cz and Rz) is connected across the unused secondary winding of the transformer. The output capacitor C12 is shown as a 'perfect' capacitor, but with its ESR in series. A 47µF cheap electrolytic will work. Alternatively, a high-cost MnO_2 solid-aluminium capacitor will last forever. If you want the over-engineered solution for lowest noise, use a 470µF low-ESR polymer solid-aluminium type with a *real* series resistance of 0.56Ω (see Fig.9). Cheap 'rubber-bung' electrolytic capacitors have short lives. The expensive capacitors are only required if a 20-year life for the power supply is expected.

Construction

Only one rail is constructed on the board, so the Zobel components can easily be fitted with a bit of jiggery-pokery in the vacated space and pads shown in Fig.6. The full overlay is shown in Fig.7. Note the earth link across pad A to what was the positive rail. This is also shown in a photo of the board in Fig.8.

Putting resistors in series with capacitors can be tricky with radial types, as shown in Fig.9. However, it's easy with a vertically mounted axial component which then becomes a 'composite radial' component, dropping into the existing pads. This is shown in Fig.10. Note that Cz is quite tall, so it's a good idea to insulate the vertical wires with sleeving, as illustrated in Fig.11. This is good practice with all vertically mounted axial components to prevent short circuits. Remember to insulate all mains wiring, including the tracks under the board, to avoid shocks.

Parts list

Note: many component positions on the PCB are unused, so the part numbering is *not* consecutive. The PCB and transformer are available from the *PE PCB Sevice*.

Resistors (All 0.25W except R1)

R1	1kΩ 0.5W
R2	750Ω
R3	120Ω
Rsurge	10Ω
Rz	2.2Ω

Capacitors

C1	10nF ceramic X7R 5mm
C2-5	100nF ceramic X7R 5mm
C10	1500µF or 2200µF 25V radial electrolytic
C11*	15µF 20V axial tantalum in series with Rsurge (or 100uF 16V electrolytic)

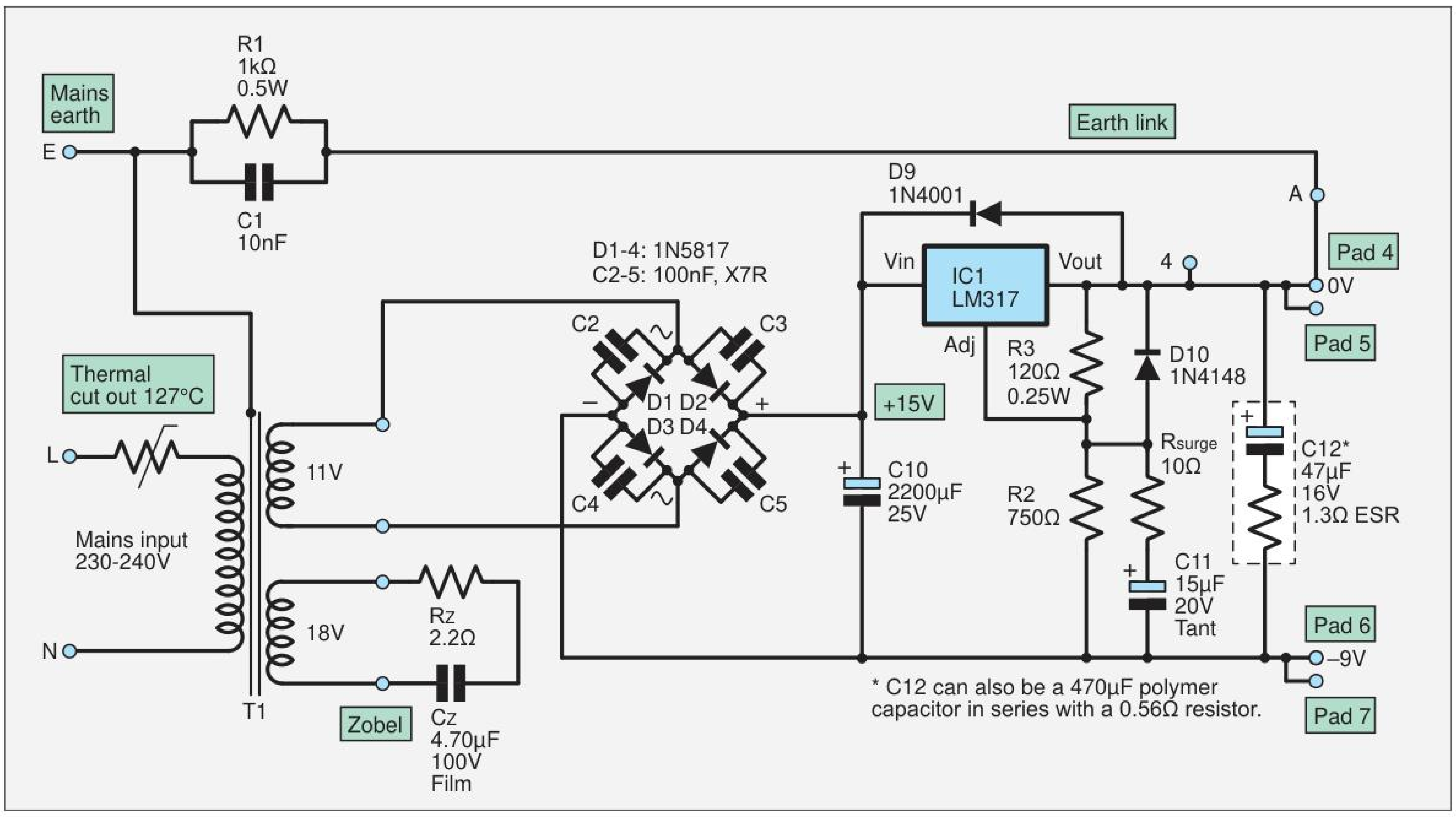
Mains earth
E
R1
1kΩ
0.5W
C1
10nF
Earth link
D9
1N4001
A
D1-4: 1N5817
C2-5: 100nF, X7R
Vin
IC1
LM317
Vout
Adj
4
Pad 4
0V
Pad 5
C2
C3
C4
C5
D1 D2
D3 D4
+15V
R3
120Ω
0.25W
D10
1N4148
Rsurge
10Ω
C12*
47µF
16V
1.3Ω ESR
C10
2200µF
25V
R2
750Ω
C11
15µF
20V
Tant
Pad 6
–9V
Pad 7
Thermal cut out 127°C
L
N
Mains input
230-240V
11V
18V
T1
Rz
2.2Ω
Cz
4.70µF
100V
Film
Zobel
* C12 can also be a 470µF polymer capacitor in series with a 0.56Ω resistor.

Fig.5. Full power supply circuit – note the resistor in series with the output capacitor (C12). This resistance can be a parasitic (within the component) or added externally.

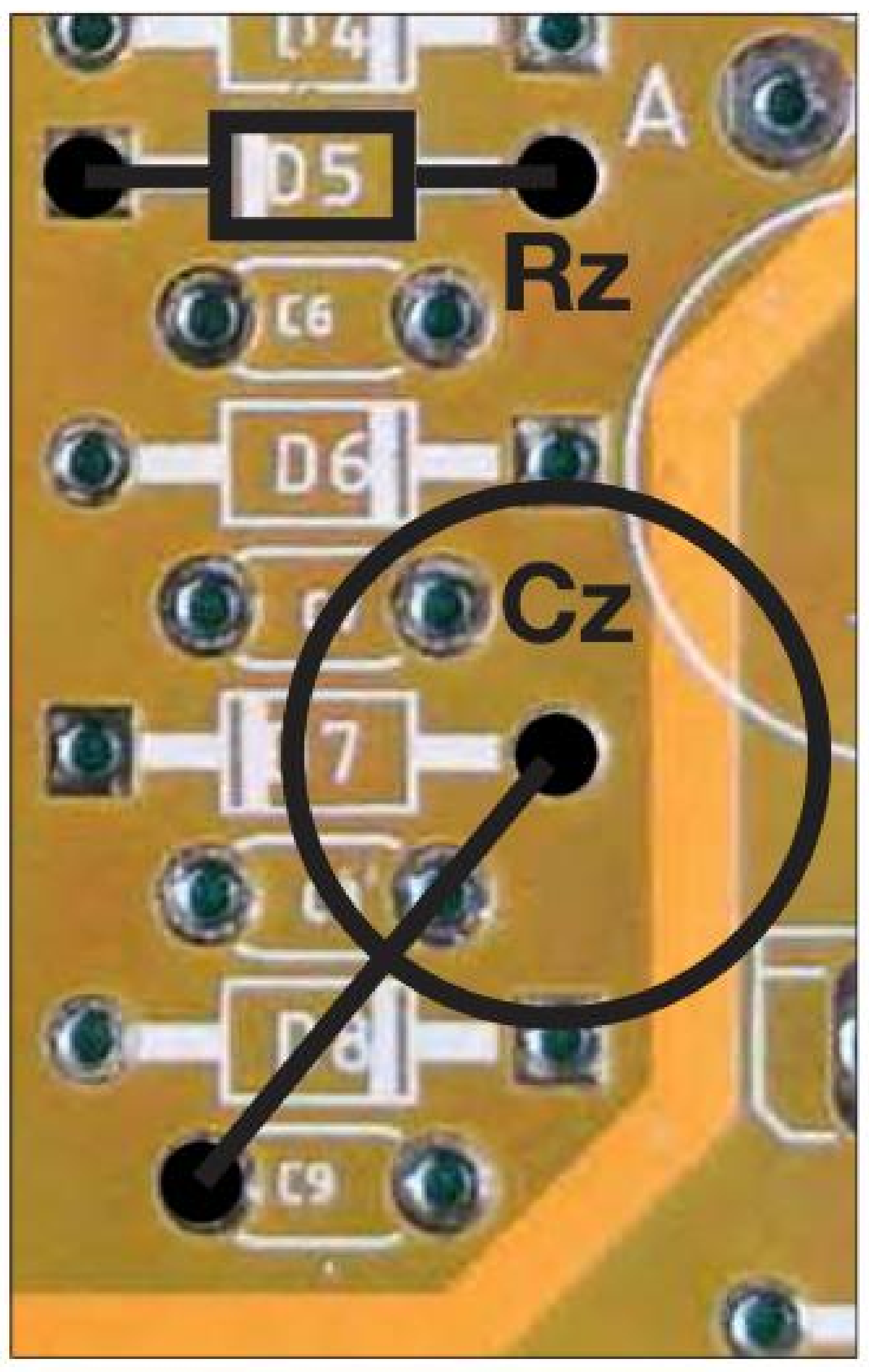
D4
A
D5
Rz
C6
D6
Cz
D7
C9

Fig.6. The Zobel network is mounted in the holes shown. Rz is placed in D5's position and Cz goes from the left-hand hole of C9 to the right-hand hole of D7.

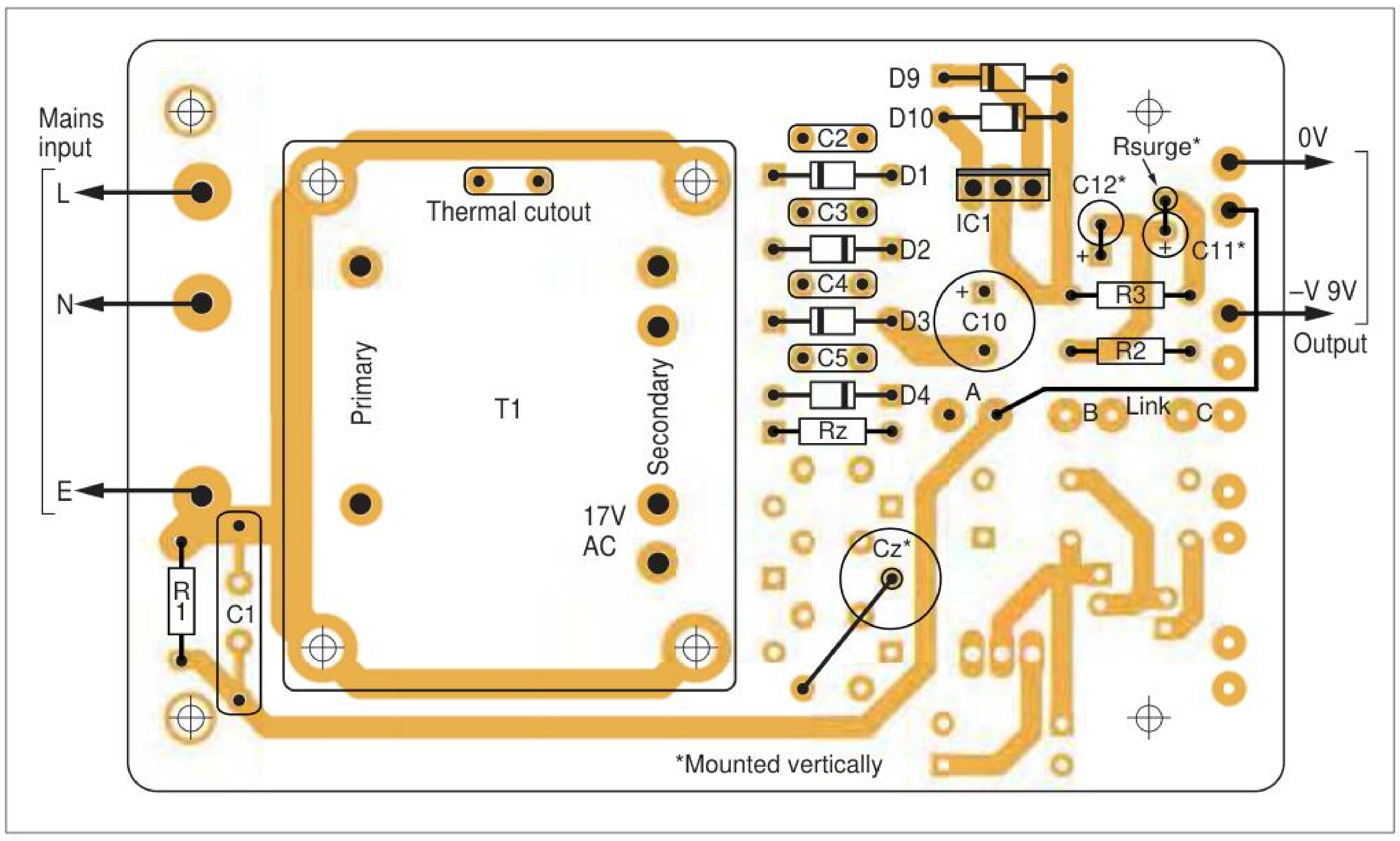

Fig.7. Overlay of the positive-earth power supply.

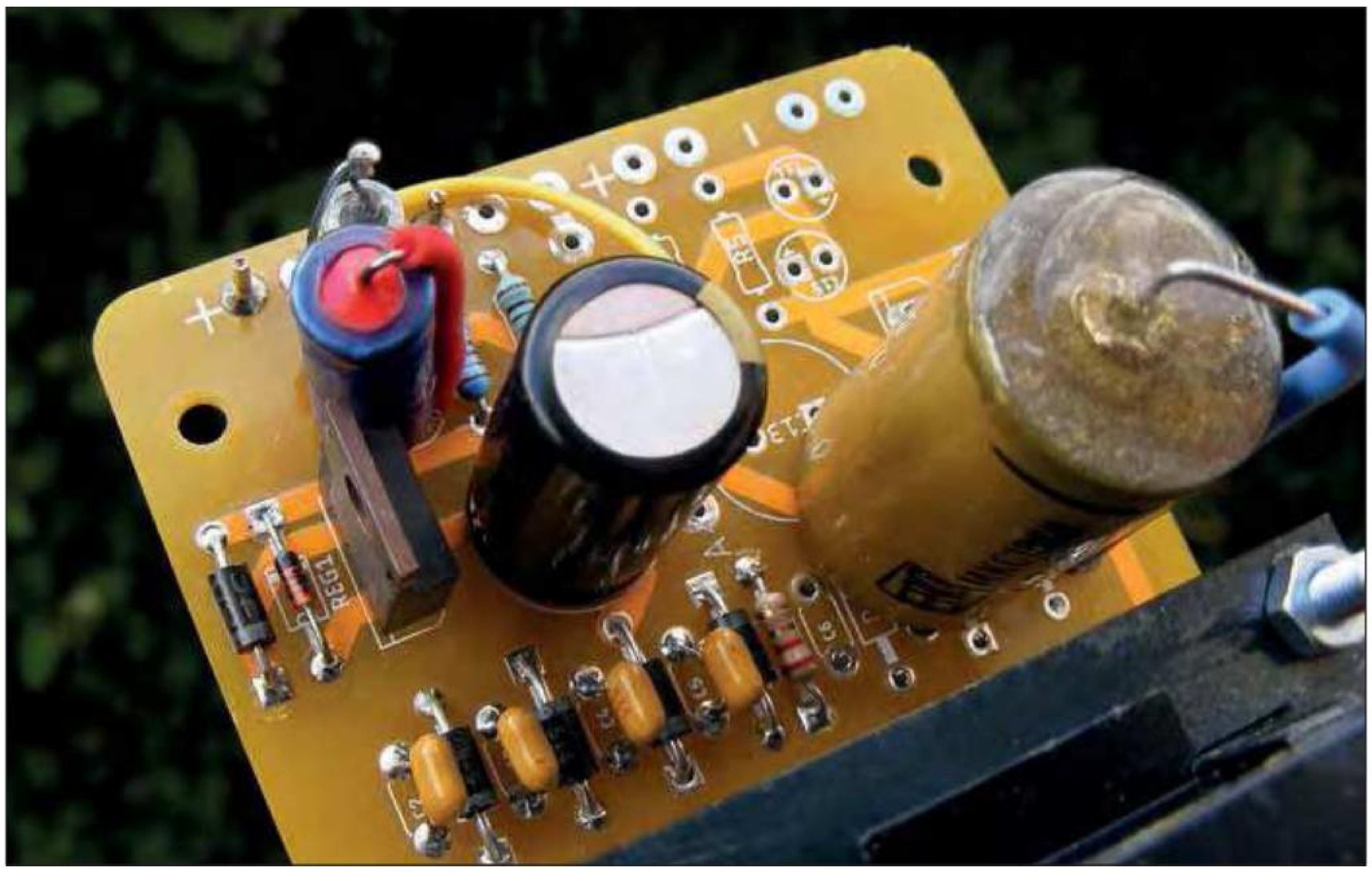

Fig.8. The completed board – view with Fig.11 to get the whole picture.

C12*	47µF 16V 123 series solid aluminium (or 100µF 16V electrolytic)
Cz*	4.7µF 160V axial polycarbonate or other non-polarised capacitor up to 10µF

*Available from me: 01597 829102
jacob.rothman@hotmail.com

Semiconductors

D1-4	1N5817 Schottky diodes 20V 1A
D9	1N4001
D10	1N4148
IC1	LM317T 1.5A voltage regulator

Miscellaneous

- Transformer: *see* last month (or use a 10 to 12V dual 200mA 6VA equivalent)
- 1mm sleeving for vertical components. Silicone is best because it doesn't melt with soldering.
- Thermal cut-out (if plastic case used) not essential for a die-cast metal box. 250mA time-delay fuse plus holder.
- 3-pin IEC mains connector
- DC power connector 2.1mm. (See *Techno Talk* last month for a detailed description of these connectors).

Installation

If putting the power supply in an old radio, make sure it is away from the ferrite rod aerial and volume control to avoid hum pick-up. I just wire PP9 battery connectors to the power supply, so it can simply take the place of the battery and be unclipped if needed. This is a good idea if the equipment is highly collectable.

Unlucky C13

I'm afraid a gremlin crept into last months PCB overlays: Figs 23, 25 and 28. There are two C13s. The smaller one is C11.

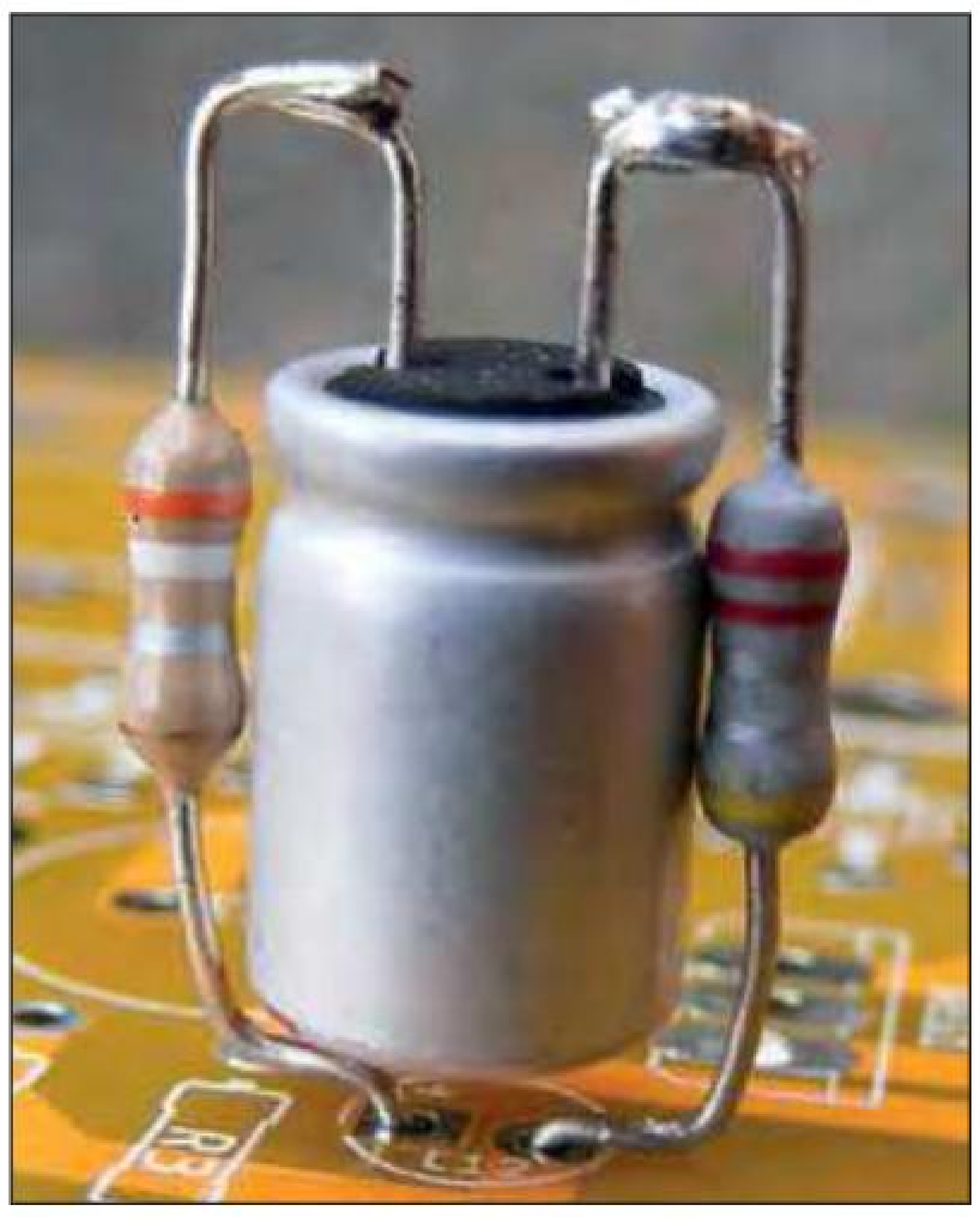

Fig.9. Putting components in series for axial components can be tricky. Here's one trick I use to add extra series resistance to solid polymer capacitors (such as C12) for damping. In this case, the resistance is split into two resistors, 0.39Ω and 0.22Ω, to give a total of 0.61Ω. The capacitor's actual ESR of 0.01Ω is just too good!

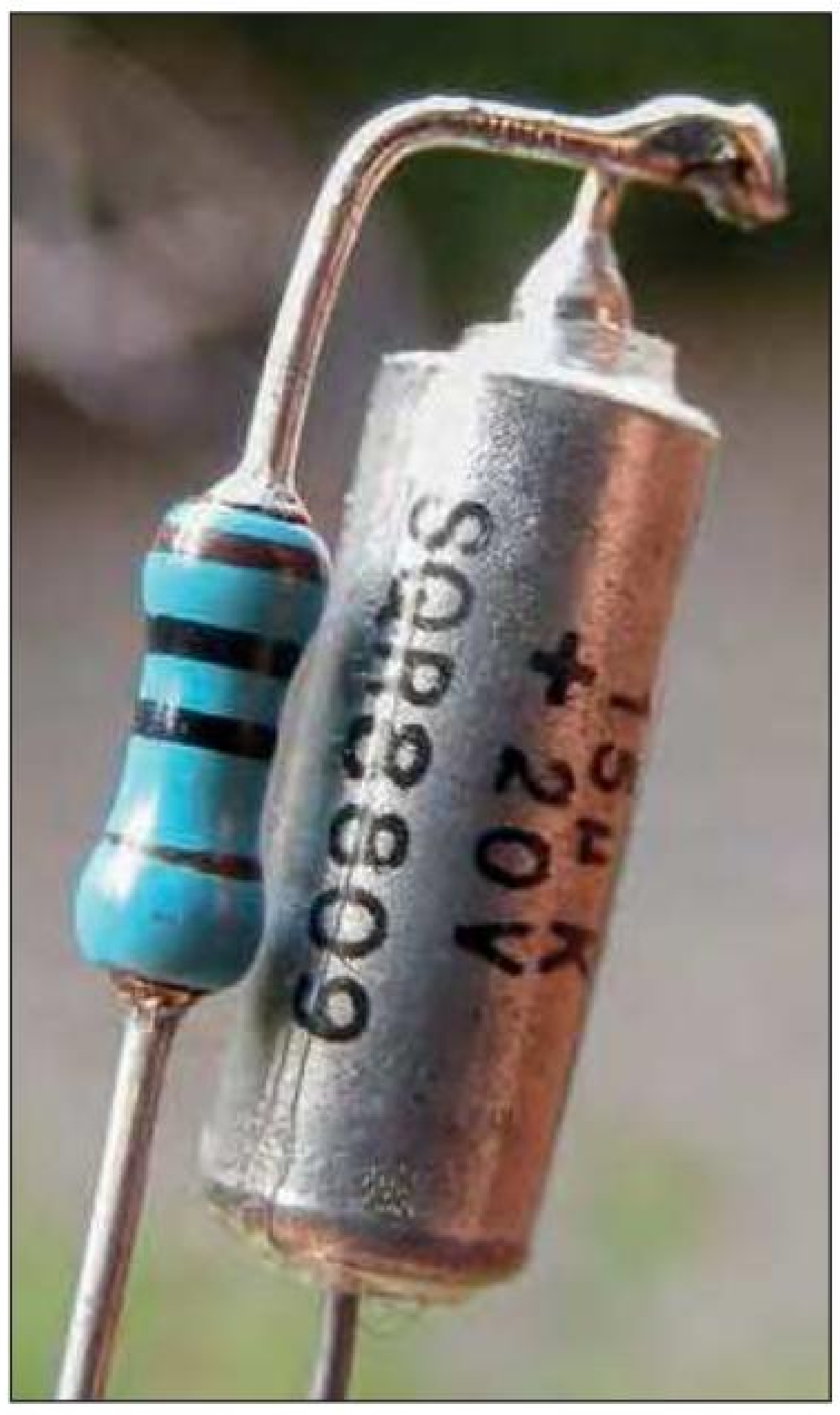

Fig.10. It's simple to put two axial components in series to make a 'composite radial' component. This is a tantalum capacitor in series with a surge protection resistor, as used for C11.

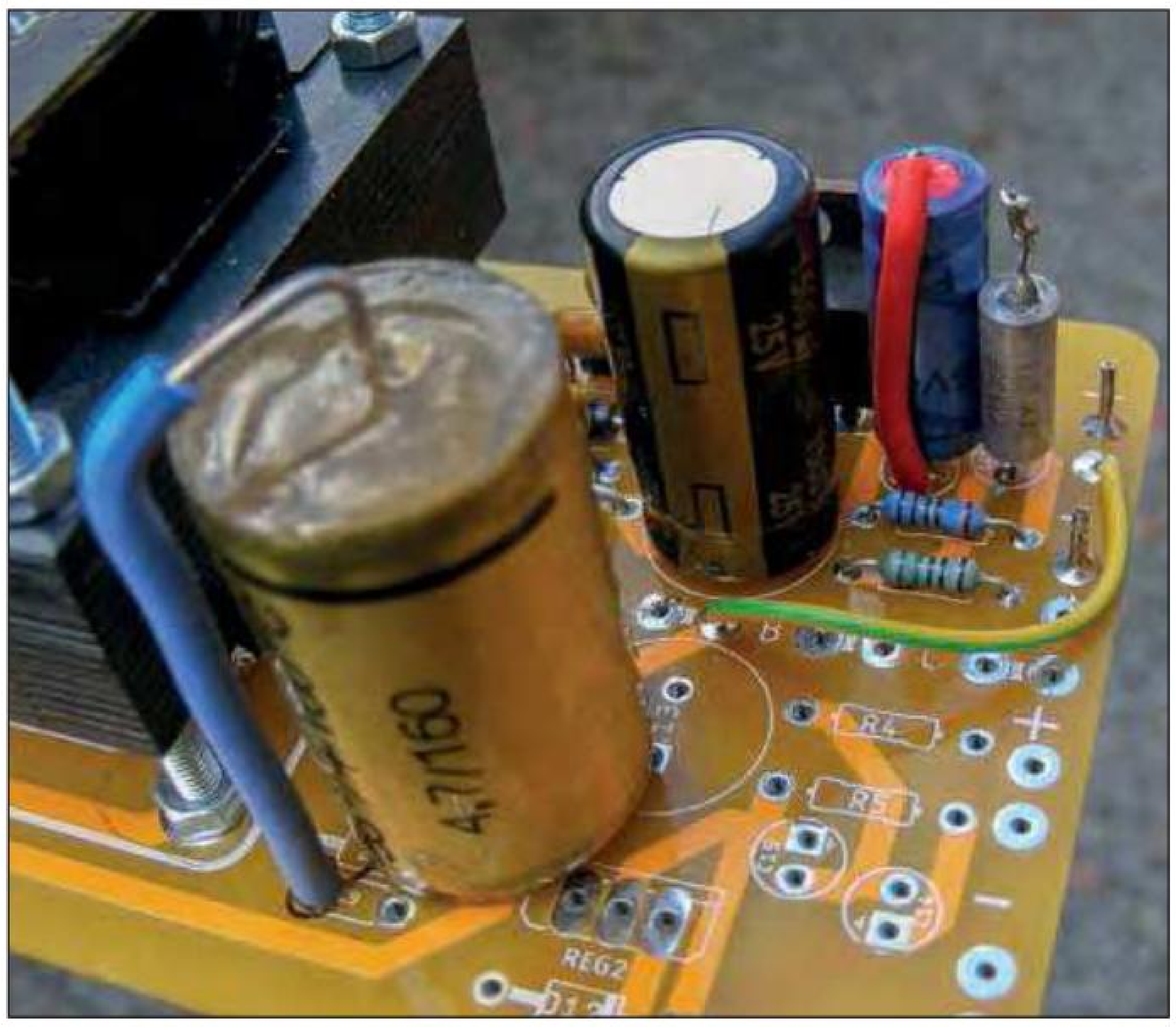

Fig.11. I used to avoid vertical mounting due to track breakage in the days of single-sided boards, but plated-through-hole boards are capable of withstanding the mechanical stress. It's good practice though to insulate the exposed vertical wires with sleeving to prevent shorts. Note the green/yellow earth reference link from point A (ground-lifted mains earth) to the old positive output terminal.

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Next Month – in the November issue

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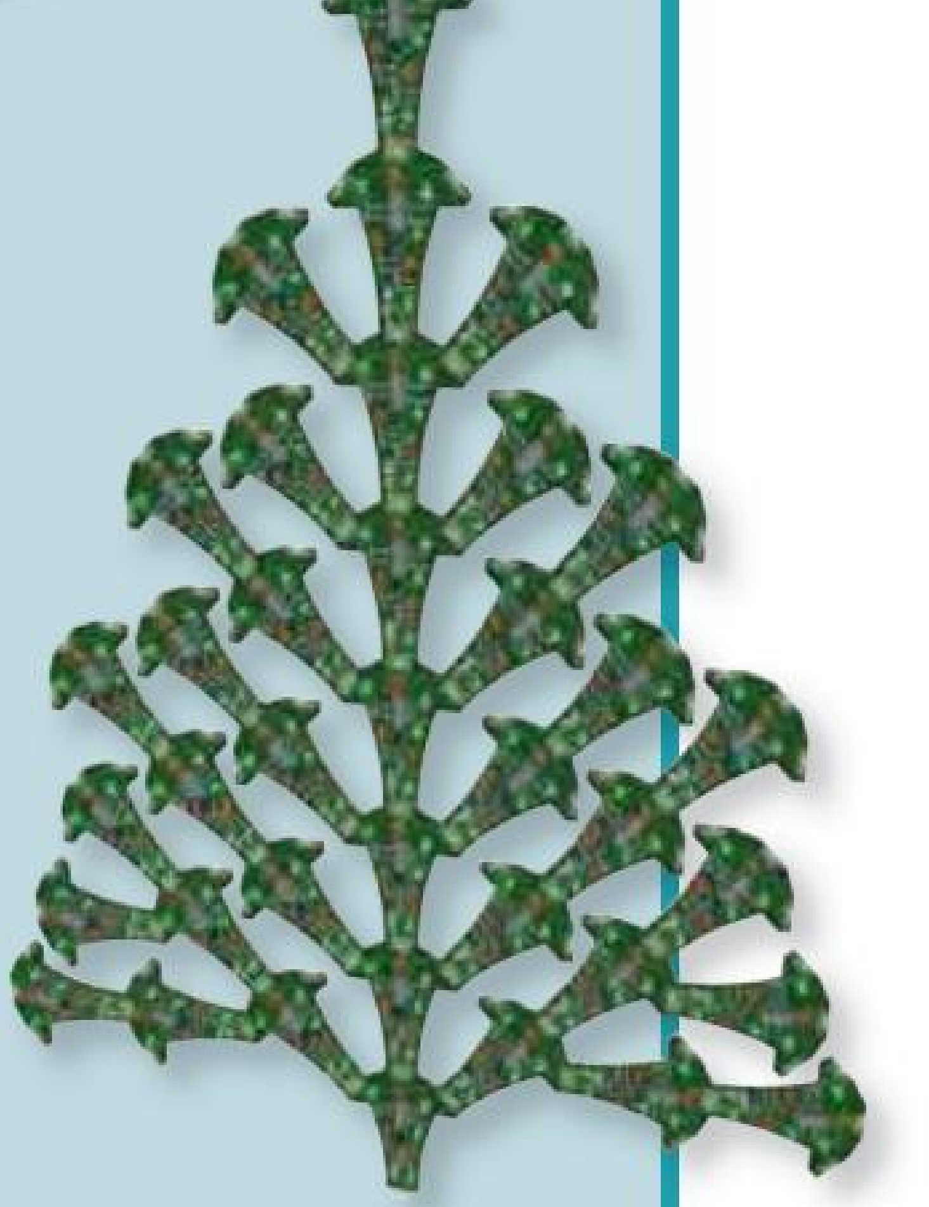
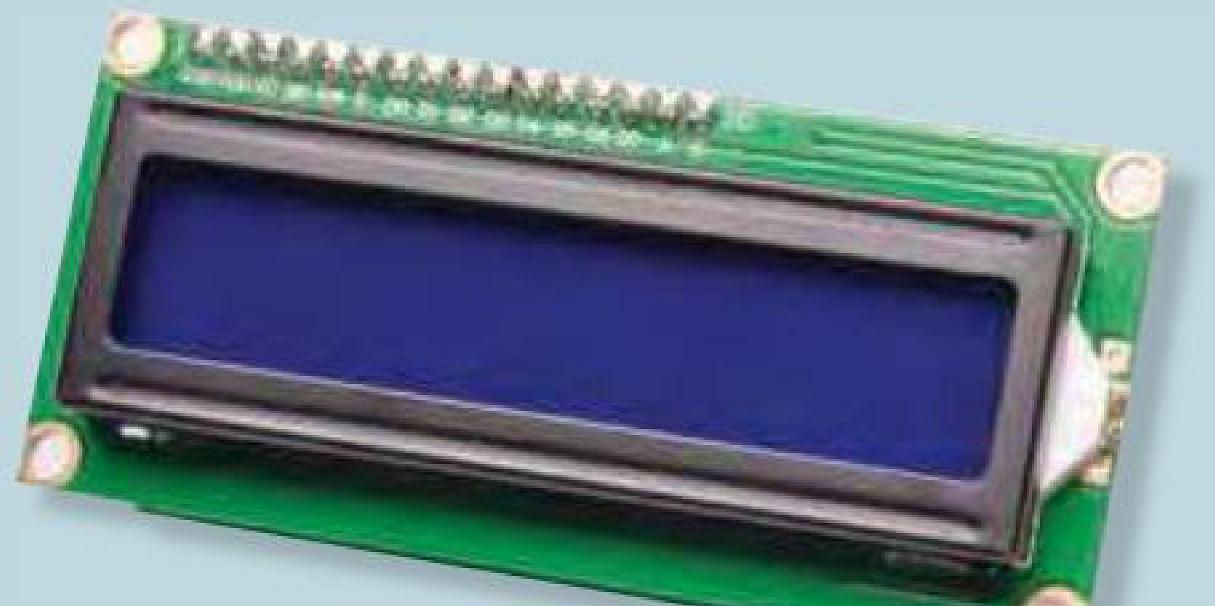


High Power – 45V/8A Variable Linear Supply – Part 2
In the October issue we introduced our new *Linear Bench Supply*. It's based around a 500VA toroidal transformer, a PCB control module fitted to a finned heatsink and two thermally controlled fans to keep it cool. Next month, we'll cover the assembly and testing details of the PCB module.



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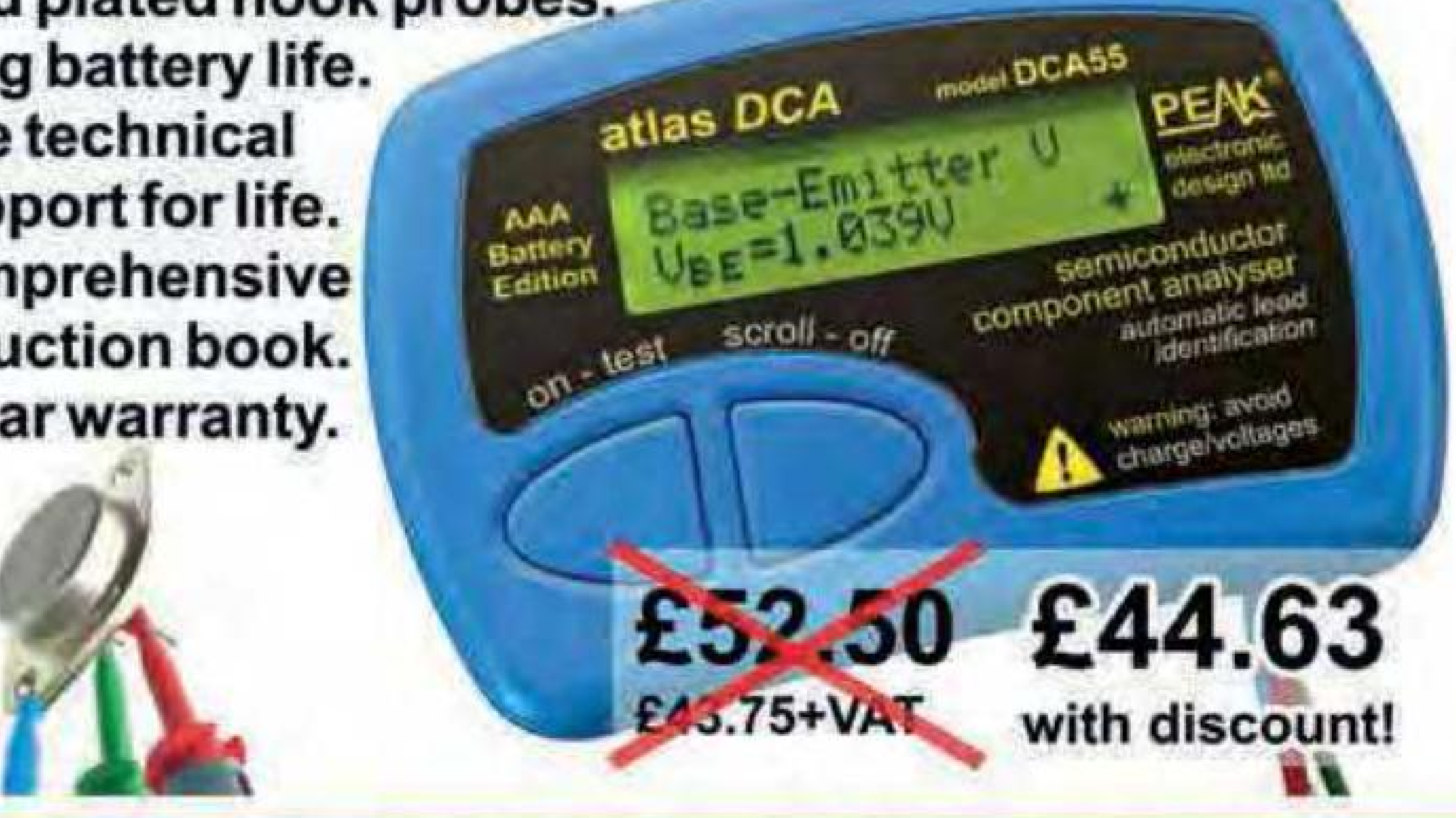
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